Practical Television Circuits

Practical Television Circuits

NEWNES

PRACTICAL TELEVISION CIRCUITS

Revised by

R. E. F. Street

LONDON
GEORGE NEWNES LTD
TOWER HOUSE, SOUTHAMPTON STREET, W.C.2

First Published 1954
Revised Edition 1968

Made and Printed in Great Britain by
Morrison & Gibb Ltd.
London and Edinburgh

Preface

In compiling this book, my aim was to include items of interest to both television enthusiasts and electronics enthusiasts. Some of the equipment described, such as the pre-amplifiers and aerials, is designed purely for television purposes, but many of the items of test-gear (the two oscilloscopes, for example) have uses throughout the field of electronics.

The details of the "Olympic II" receiver which are given here include the modifications and amendments suggested by the designer since the publication of the original constructional information. The opportunity has also been taken of bringing up to date the other circuits presented, and I should like to emphasise that this edition of *Practical Television Circuits* is completely new, containing nothing from previous editions.

I must express my thanks to Mr. A. T. Collins (Managing Editor of Newnes' "Practical Group" of publications) for the assistance he afforded me in the initial stages and during the final preparations for printing. I am indebted to him for permission to use material from *Practical Television.*

R. E. F. STREET

Derby,
September 1967

Contents

Section 1

PRE-AMPLIFIERS

Chapter One

A NUVISTOR PRE-AMP FOR BAND III

ALTHOUGH there are many "booster" transmitters, areas remain where television (especially on Band III) is of poor entertainment value owing to weak signals, resulting in noise on both picture and sound. The obvious way to improve results in such areas is to pre-amplify the signal before it is fed into the receiver. However, a limit is set in the application of this method by the noise in the pre-amplifier (the current flowing in the valve or transistor concerned is not constant but continually varying, and gives rise to an output signal). Normally, this noise is insignificant compared with the signal being amplified, and is of no importance. In fringe areas, though, the signal received may be of the same order as the noise from the pre-amplifier. Thus, a limit is set to the useful amplification which may be obtained; the signal received must be sufficiently strong to override the noise in the pre-amp.

It is an obvious advantage if the noise introduced by the pre-amp is at as low a level as possible, and it is true that modern transistors are better than valves for use in pre-amps. Nevertheless, valved circuits are still capable of giving good results.

This valved pre-amp is designed for Band III operation, and uses a 6CW4 Nuvistor triode in a single-valve circuit. The 6CW4 has a metal envelope with a ceramic base on which all the internal electrodes are supported. The small size (about $\frac{3}{8}$in $\times$ $\frac{3}{4}$in high) enables a small, neat and efficient layout to be employed.

Circuit

The simplicity of the circuit used should be apparent from Fig. 1. Power requirements are 0·15A at 6·3V and 8mA (h.t.) at 70V. A conventional power supply may be used if the resistance R3 is calculated to give 70V at the anode of the 6CW4. (In the prototype, with 180V h.t., R3 needed to be 8·2kΩ).

As might be expected, the circuit needs neutralisation, and capacitive, rather than inductive neutralisation is provided by TC1 and TC2. The components used are not critical, but TC1 and TC2 must have low minimum values. Effective neutralisation cannot be obtained with 3-30pF concentric trimmers, and types having a minimum capacity of 1pF or less *must* be employed. If the minimum capacity is greater than about 1pF, the pre-amplifier will be unstable and it will be impossible to obtain good results.

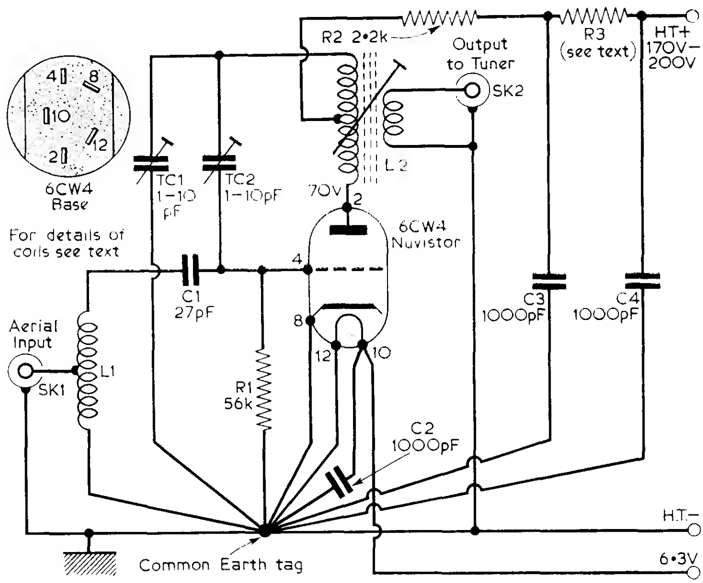


Fig. 1. The circuit diagram.

Construction

The prototype was connected using a small cough-lozenge tin as a chassis, this being used as it was quite large enough to accommodate all the components and the tin-plated surface enabled components, including the valve-holder, to be soldered directly to the chassis. The thin metal also enabled all necessary holes to be made using a sharp-pointed instrument and, thus, no drilling was necessary.

The layout and wiring given in Figs. 2 and 3 should be followed exactly, particularly the wiring of the heater circuit. The heater leads are not brought through the chassis in the positions shown for convenience, but to avoid decoupling troubles. Note that the "earthy" side of the heater is taken to the common soldering tag, and pin 12 of the valve is also earthed to this tag. The heater wire should not be taken to pin 12 and from there to earth. Likewise, all components shown wired to the common earth tag in Fig. 1 should be wired separately to the tag as shown in Fig. 2. This procedure prevents certain impedances from being made common to two or more circuits.

At first, it may not be apparent that C3 and C4 are wired to the common earth tag. In fact, they are soldered above the chassis next to the bolt which is used to hold the common earth tag to the underside of the chassis; see Fig. 3. If desired, the common earth tag may be soldered to the chassis for extra rigidity.

Adjustment

If it is possible to obtain a Band III picture without the use of a pre-amp, the receiver should be set to give the best picture possible. If a Band III

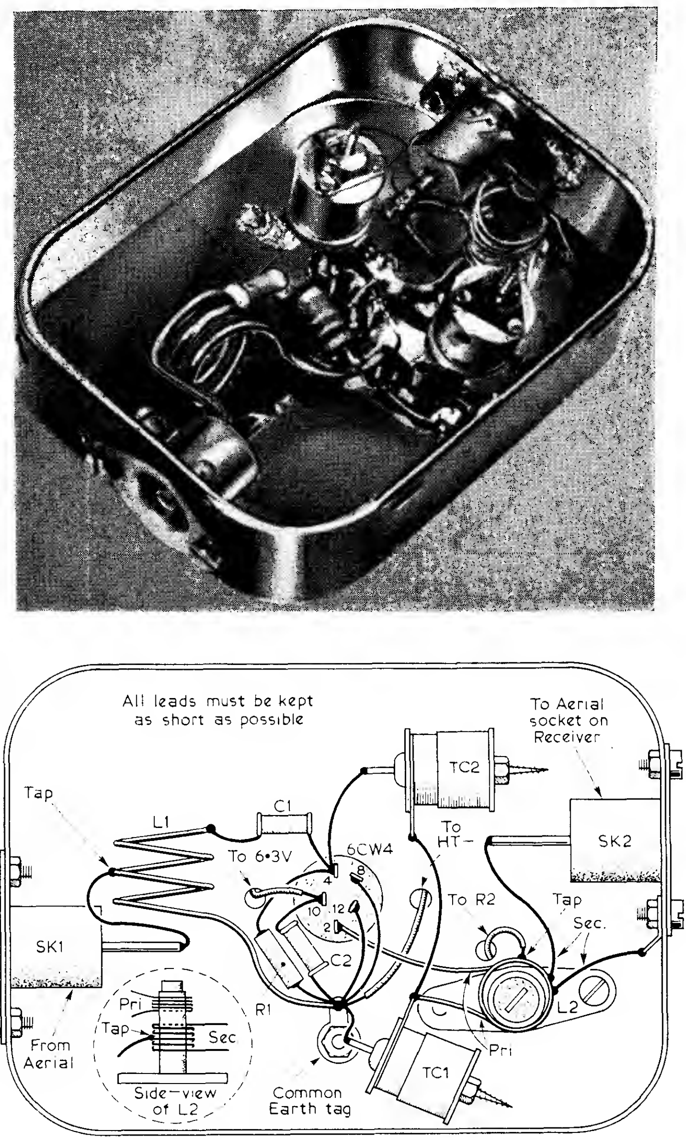
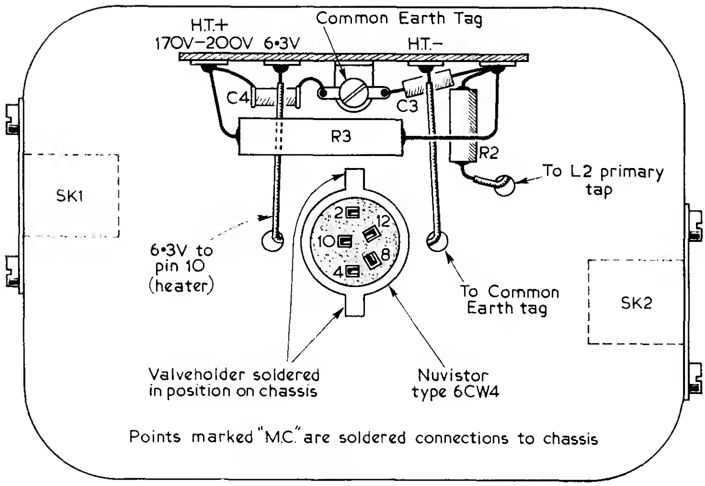


Fig. 2. The under-chassis view and wiring diagram.



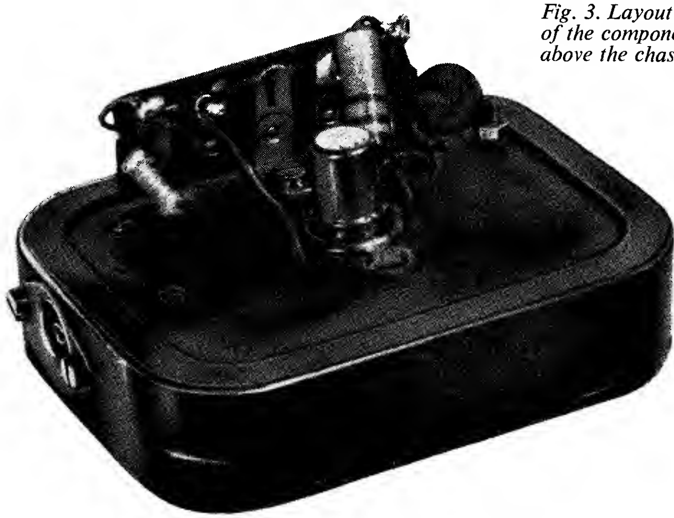


Fig. 3. Layout and view of the components above the chassis.

picture cannot be obtained, the TV receiver should be set to the correct channel number and the fine tuner control turned to the midway position.

The aerial lead is then removed from the set and plugged into the input socket of the Nuvistor pre-amp. A short lead is then prepared with a coaxial plug on each end; this is used to link the output of the pre-amp to the aerial socket of the receiver.

At first, the aerial input tap in the pre-amp (on L1) is set to the centre of L1. The h.t. feed from R2 to L2 is soldered initially to the centre point of

this coil. The trimmer TC2 is set to minimum capacity and TC1 to about half capacity—say 5pF. The unit is then linked to the set using the prepared cable.

Upon switching on, it may be found that results are as they were before or even worse. Screw in the core of L2 for best results. Once some indication of the signal has been seen or heard, the core of L2, TC1 and TC2 must be adjusted to obtain optimum neutralisation. Alteration of the setting of the core of L2 will mean that the setting of TC1 must be altered to keep L2 at resonance—if the core is moved into the coil, the capacity of TC1 must be decreased and vice versa.

After each change in the setting of the core of L2, and TC1, TC2 is increased in capacity until any further increase results in instability—shown when the screen of the receiver becomes uniformly bright and a hum is heard from the loudspeaker. The optimum settings for the core of L2, TC1 and TC2 will soon be found, although perseverance may be necessary if the signal is weak.

Adjusting L1

When results may be improved no further, the tapping point of the aerial input on L1 may be altered in an endeavour to secure a better signal. Coil L1 may be tuned to the signal by checking whether it is at resonance by holding a dust core near to it; if results improve, then the inductance of L1 needs to be increased and the turns of L1 should be squeezed together. If results deteriorate when the dust core is held near to L1, then the turns will need to be spaced further apart.

It will be noted from Fig. 2 that L2 is wound with a larger diameter than its former (which serves to hold the dust core rather than to support the coil) and the h.t. tap may easily be moved along the coil (soldering it each time). Altering the tap may mean altering the settings of the core of L2 and TC1 and TC2 again for best results.

It cannot be emphasised too greatly that in areas of poor signal level, all these adjustments are quite critical for optimum results to be obtained; and they are, to some extent, interdependent, although adjustments in the input circuit have less effect than those in the output circuit. However, alterations to the setting of the core of L2 *must* be followed by alterations to TC1 and TC2.

The adjustment procedure is long to set out, but is quite easy once initial adjustments have been tried.

Curing Mis-Match

Note that if there is a mis-match between the receiver input and the output of the pre-amp, then the length of the cable linking them may influence results—try increasing or decreasing its length. It is also worth remembering that low-loss downlead is essential—in areas of weak signal, more than half of the received signal may be lost between the aerial and the receiver.

List of Components

Resistors:
- R1 56kΩ ½W
- R2 2·2kΩ ½W
- R3 about 8·2kΩ 1W (see text)

Capacitors (ceramic):
- C1 27pF
- C2 1000pF
- C3 1000pF
- C4 1000pF

TC1, TC2 1–10pF concentric trimmers, with a minimum capacitance of 1pF or less

SK1, 2 Coaxial sockets

6CW4 Nuvistor triode valve

Coil former and dust core (¼in)

Details of Coils

L1 3 turns of tinned (or silver-plated) copper wire (the *single-stranded* wire sold as "earth" wire is very suitable). The turns are spaced by the diameter of the wire and the coil is wound on a ⅜in former which is then removed. One end of the coil is left long for soldering to the common earth tag. The aerial tapping on to L1 is arranged to give optimum results during the alignment of the unit.

L2 Primary—4½ turns of tinned copper wire (about 22 s.w.g.)
Secondary—3 or 4 turns of d.c.c. copper wire (about 34 s.w.g.) positioned at the TC1/TC2 end of the primary.

Note: As mentioned in the text, L2 is wound with a diameter of about $\frac{5}{16}$in—larger than that of the former used which serves to accommodate the dust-core, rather than to support the coil. This is to enable the h.t. feed easily to be tapped on to L2 at the position.

The number of turns on the coils may have to be altered to secure best results, but the above data should be used for initial experiments.

Chapter Two

A HIGH GAIN BAND III PRE-AMP

A MEANS of improving reception is always welcomed by the experimenter, professional or amateur. This Band III pre-amplifier is developed around the alloy-diffused transistor type AFZ12, by Mullard, which is particularly suited to working at Band III frequencies. When used in the circuits described, it is capable of high gain at a low noise level. This last-mentioned characteristic is very important if a good picture is to be obtained.

It should be pointed out that constructors are advised to use the correct transistor for this pre-amplifier. It is tempting to use cheap substitutes, but at 200Mc/s, only a first-class transistor will give satisfactory results.

Two Circuits

It will be seen that two circuits are shown for the pre-amplifier; one uses the common-base mode (Fig. 1), and the other the common-emitter mode (Fig. 2). The common-base circuit gives slightly higher gain but has a higher noise figure—5·5dB—compared with 3·9dB for the common-emitter circuit. The reason for the difference in noise figures stems from the fact that better matching is possible with the common-emitter circuit.

The average gains for the common-base and common-emitter configurations are 12·5dB and 11·0dB respectively, for a nominal emitter current of 1mA. The bandwidths for the pre-amplifiers are 2·7Mc/s (common-base) and 5·0Mc/s (common-emitter).

It should be noted that these figures were taken at 200Mc/s; higher gain can be expected at the lower frequencies of a number of the Band III transmissions.

The pre-amplifier is perfectly stable provided certain precautions are taken, i.e. the output must be screened from the input, which is the reason for the design adopted. It will be noted that the power supplies are decoupled in the interests of stability.

Construction is straightforward and should be carried out in fairly stout copper sheet where possible. The main chassis should be marked out in the flat state and drilled and then formed over a block of wood which has been cut to the inside dimensions of the base of the chassis. Some constructors may complete the chassis by fitting sides and a lid which can be soldered in position after final adjustments have been made. The amplifier would then be protected from damage and dust.

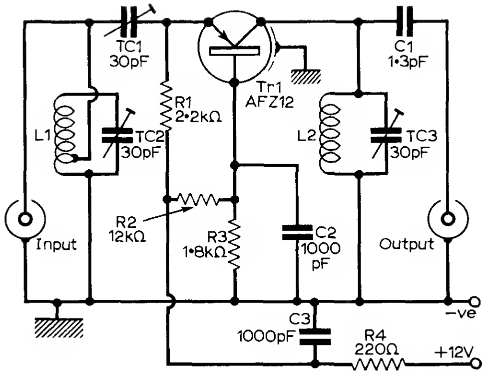


Fig. 1. The AFZ12 transistor used in the common-base mode as a pre-amplifier.

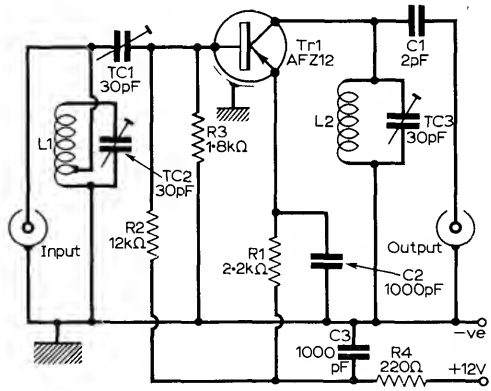


Fig. 2. The common-emitter circuit used in the pre-amplifier.

Transistors must always be protected against heat when soldering is carried out, so in the interests of safety, where one or more components have a common connection with the transistor, these components should *first* be soldered together, leaving a single joint to be made to the transistor. The last joint should be made using a heat shunt on the transistor lead.

Coils

The coils are wound and formed so that a "foot" is made to facilitate soldering (see Fig. 3). The coils are self-supporting and should be wound with the diameter of the wire between turns. Beehive-type trimmers are used for tuning and to ensure that they are mechanically rigid when soldered, the stem should be pressed into a B.A. nut to form a base which is soldered to stem and chassis. Fig. 3 shows the common-emitter wiring.

The transistor must be carefully positioned so that the screen passes between output and input. Short lengths of sleeving should be fitted to transistor leads to obviate short-circuits. The transistor holder should be

carefully fitted to hold the transistor firmly under normal circumstances (Fig. 4).

A refinement would be a simple wooden case made to fit amplifier and supply-battery with on/off switch.

Tuning

Having constructed the amplifier and checked against wrong connections and unwanted solder, tuning can be proceeded with. Starting with TC3, this should be adjusted for maximum response. TC2 should next be adjusted for maximum and lastly TC1. This procedure should be repeated until no further improvement can be obtained, after which the beehive trimmers should be sealed with wax to prevent movement.

If it is found during tuning that maximum output cannot be obtained within the tuning range of TC2 or TC3 then L1 and L2 can be adjusted by

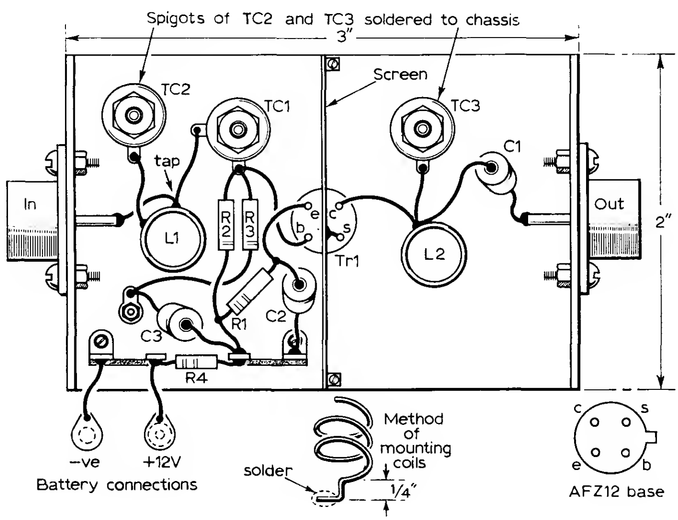


Fig. 3 (above). The wiring diagram of the common-emitter circuit.

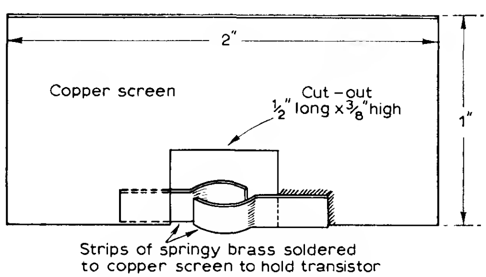


Fig. 4 (right). Details of the copper screen and transistor-clip.

separating or squeezing the turns slightly. Separating the turns increases the operating frequency and squeezing them together lowers it. If either TC2 or TC3 is at minimum capacity (open) for maximum results, then the associated coil requires *slightly* stretching until the trimmer will tune the signal in and out.

If these instructions are carefully carried out maximum results will be obtained. It is important that first-class components be used throughout if the best performance is to be gained.

Choice of Circuit

In conclusion, then, the experimenter living in an area of very weak signal, together with local interference, would be best advised to use the common-emitter circuit with its low noise level, whilst for stronger signal areas the common-base model would be most satisfactory. If maximum gain is, however, required and the constructor is prepared to accept a slightly noisier background, then the common-base configuration is best suited.

List of Components

Resistors:
- R1 2·2kΩ
- R2 12kΩ
- R3 1·8kΩ
- R4 220Ω
- All 10% ¼W carbon

Capacitors:
- C1 1·3pF (Fig. 1), 2pF (Fig. 2)
- C2 1000pF ceramic
- C3 1000pF ceramic
- TC1, TC2, TC3 3pF–30pF concentric trimmers

Transistor:
- Tr1 AFZ12 (Mullard)

Miscellaneous:
- 3 single tag-strips, 2 coaxial sockets, sheet copper for chassis and screen (18–22 s.w.g.) connecting wire, nuts and bolts etc.

Details of Coils

- L1 2 turns, tapped at ½ turn from chassis.
- L2 2½ turns

Both coils are wound to an outside diameter of ⅜in with 18 s.w.g. tinned copper wire.

Chapter Three

A TRANSISTORISED PRE-AMP

In some locations, which are cut off from the signal by a local hill, it is often necessary to mount the aerial towards the top of the hill and feed the signal to the set through coaxial cable. This invariably results in far better signal pick-up at the aerial, but the extra long downlead often dissipates most of the signal thus gained, due to its attenuation (losses), and the net gain is almost zero.

Shared aerials are sometimes employed by a group of viewers located in a valley. The cost of the elaborate hill-sited aerial system is shared by the viewers and an amplifier (or set of amplifiers) is used to counter the losses of the downlead and the losses involved in tapping the signal to the various receivers.

There are many commercial valve-type amplifiers (repeaters) already available for this kind of application, and transistorised amplifiers of a similar kind have also been developed. There are also many hundreds of shared-aerial systems working quite well in the hilly parts of the country.

Mast-head Amplification

Where the signal is weak to start with, the amplifier should be situated as close as possible to the aerial, and the signal should be fed down the extra long downlead at high level. If, even after mast-head amplification, the signal at the end of the downlead is insufficiently strong for distribution or for operating a receiver, then a second amplifier should be used at the receiver (see Fig. 1).

Amplification close to the aerial ensures that the best possible signal-to-noise ratio is secured, and subsequent distribution or amplification does not detract from this too much. However, if the aerial signal were allowed to be severely attenuated by the downlead before initial amplification, then the signal-to-noise ratio would be very much worse.

Power Supply Problems

Mast-head amplification means that a power supply must be conveyed to the hill-aerial site. This can be expensive if a mains system is installed, but is a problem which is often solved by feeding a low voltage power supply from the receiver end to the remote mast-head via the signal-carrying coaxial

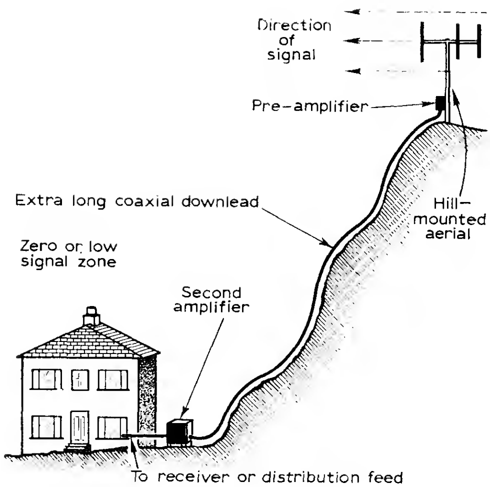


Fig. 1. An extended downlead (as shown) requires mast-head amplification, and a further local pre-amplifier may be required if the downlead is long.

downlead. This is difficult where valve equipment is used, owing to the fairly high power requirements, but it is possible.

Low Current

The problem is simplified with transistorised equipment, of course, since then only a few milliamperes of direct current are required to power the mast-head amplifier.

The transistor amplifier which is about to be described was developed for use either close to the receiver (for ordinary pre-amplification) or for use on a shared-aerial system, where the amplifier is likely to be mounted and powered at a point some 200yd or 300yd remote from the receiver, close to a hill-sited aerial system. Facilities will also be described whereby the power for the pre-amplifier can be fed through the coaxial downlead.

Circuit Description

The complete circuit of the pre-amplifier is given in Fig. 2. Two transistors are used, Tr1 and Tr2, and the circuit is reasonably conventional. Both transistor stages are of the common-base mode, the stabilising components being R4 and R5 for Tr1 and R8 and R9 for Tr2. The bases are held to "earth" at signal frequencies by C2 and C5.

The aerial signal is applied, via C1, to the emitter circuit of Tr1, across R2. Resistor R1 is included to maintain the best possible match between the aerial feeder and the amplifier. This resistor is necessary because the input impedance of Tr1 is somewhat above 75Ω, but in some cases, especially where there is a reasonable length of feeder between the aerial and the

amplifier input, the resistor can be omitted with slight improvement in overall gain.

The collector of Tr1 is loaded and coupled to the emitter of Tr2 by a form of π-network coupling, comprising L1, C3, R3 and C4, together with the collector capacitance of Tr1. A suitable impedance match for Tr2 emitter circuit is provided by the tap on L1.

A similar form of tuned circuit is used in the collector circuit of Tr2 to feed the amplified signal to the receiver or downlead and, again, a tap is used on L2 for impedance matching. Resistor R10 is necessary to provide a reasonable load impedance to the extended downlead. Where the downlead is 200yd or 300yd in length, the "buffer" effect of the cable is sufficient to stabilise the output impedance and the resistor can be omitted with some improvement in gain.

The amplifier will operate from a battery of 9V–12V, and a simple toggle switch is included in the positive supply lead for switching on and off when the amplifier is used close to the receiver.

Construction

The amplifier can be made remarkably small, and Fig. 3 gives a suggested design in a tobacco tin, of dimensions about $4\frac{1}{8}$in × 3in × $\frac{7}{8}$in deep. The actual circuitry takes up approximately half the available space, and the remainder can be used to house one of the small 9V batteries designed for transistorised equipment, thereby making the unit completely self-contained.

Fig. 3 shows the wiring and the layout of the components which are

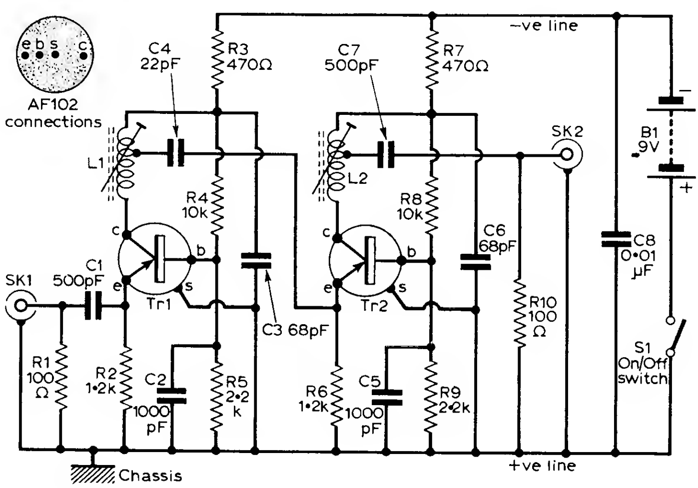


Fig. 2. The complete circuit of the pre-amplifier.

secured to the tags on three five-tag tag-strips. The two outside tags on each strip are clamped to the chassis by 6B.A. nuts and bolts. These tag-strips can be cut down from multi-tagged strips available from radio component shops. The precise design of the strip, provided it is of the miniature variety, does not really matter as long as three "isolated" tags are available on each strip. The remaining two tags are "earthed" to chassis by their fixing brackets.

Although there is no excessive rise in temperature inside the amplifier under normal operating conditions, it is as well to drill several holes in the corners of the box where the transistors are located, as shown in Fig. 3. Holes should also be drilled at the top of the box, in line with the coil formers, to allow adjustment of the tuning cores when the lid is fitted on to the tin.

There is sufficient room to accommodate resistors of the miniature $\frac{1}{2}$W, fully insulated, carbon-composition variety, while the capacitors are either

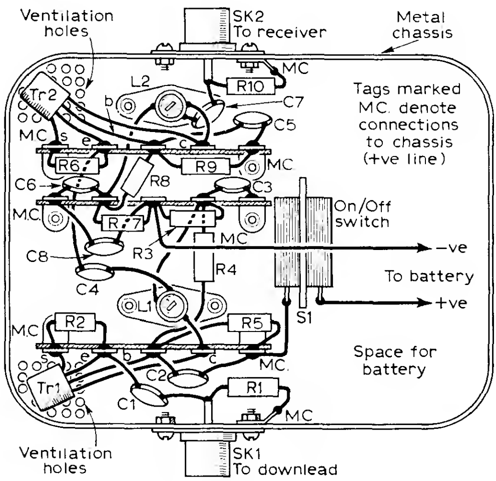


Fig. 3. The wiring diagram of the pre-amplifier (which is built in a 2oz tobacco tin).

Hi-K midget ceramics or disc ceramics (see Components List). Standard coaxial sockets are used for connecting the aerial feeder and the cable to the set or distribution line.

With the recent developments in v.h.f. transistors, there are several types which are suitable. The original design incorporated Mullard OC171's, but for Band III operation, the Mullard AF102's are recommended. Types T1832 and 2N502 by Semiconductors Limited and types 2G101 and 2G102 by Texas Instruments have also been used at Band I frequencies. It is, of course, essential to employ v.h.f. transistors. The ordinary medium-frequency and low-frequency components are no good at all for this application.

Coils

The coils are wound on polystyrene formers of height 0·87in and diameter 0·31in. The actual number of turns required to tune over a particular channel

depends very much on the construction of the unit and on the type of transistors used, and the coil winding table must be used only as a guide. In most cases, however, provided the layout has been followed fairly faithfully, the required channel will be within the range of the cores if the turns given against a particular channel in the table are adopted.

The tapping point on the coils affects both the overall gain and bandwidth of the amplifier. For television channels, with a good response to both sound and vision signals, the tapping point should be 25% along the length of the actual coil wire from the resistor end of the coil (i.e. from the end remote from the collector).

It is best to wind the coil first without the tap to find out the length of wire required *for the winding proper*, and then measure off a quarter of that length. At the appropriate point the enamel should be scraped off the wire with a pen-knife, taking extreme care to avoid cutting the wire. The tapping wire, which should be of the same gauge as the coil wire, should then be soldered to the coil wire, using the smallest possible amount of solder but avoiding a dry joint.

The coil may then be rewound correctly, and the windings may be anchored with polystyrene cement or similar low-loss adhesive.

Heat Shunts

Owing to the small length of lead-out wires from the coils, transistors and other components, an instrument-type soldering iron should be employed for the construction, and heat shunts should be clamped on the lead-out wires when performing the soldering. This applies particularly to the transistor lead-out wires.

Setting-up

Make sure that the battery is connected the correct way round, as otherwise the transistors will almost certainly be ruined. Connect the amplifier between the aerial and the receiver and, with the receiver switched to the appropriate channel, adjust the cores in L1 and L2 for the best possible sound and vision.

If tuning of any coil appears to occur with the core fully removed, a turn or so should be taken from the coil (increase the turns spacing on Bank III channels), while if the circuit only just comes into tune with the core fully in the coil, a turn or so should be added (decrease the turns spacing on Band III channels).

The power gain between the loaded sockets is of the order of 30dB on Band I channels and somewhat below this (depending on the type of transistors used) on Band III channels. Nevertheless, a very useful increase in gain is achieved with an excellent noise figure.

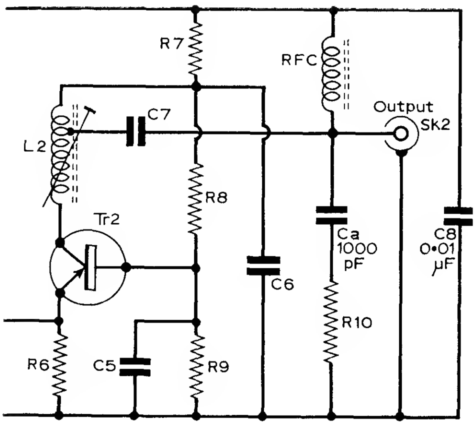


Fig. 4. The revised output circuit of the pre-amplifier for use when power is supplied via the downlead.

Remote Powering

For remote powering, the output arrangements should be altered a little as shown in Fig. 4. Here Ca has been added in series with R10 to avoid short-circuiting the supply current, and a small r.f. choke (RFC) has been introduced between the inner of Sk2 and the battery negative line.

The choke allows the unrestricted passage of battery current (from the remote end), while acting as a high impedance to the signal.

At the remote end, battery current is fed into the downlead via the system shown in Fig. 5. Signal continuity from the downlead to the receiver (or distribution system, which may include another amplifier at this point) is maintained via Sk2B, C1A and Sk1A. Power from a local battery is applied to the downlead through S2A/B and the r.f. choke (RFC), the positive connection on the battery being applied to the braid of the coaxial cable. Capacitor C1A acts as a d.c. isolation device and prevents the battery current from being short-circuited by d.c. continuity across the aerial socket of the receiver or second amplifier, while C2A acts as an r.f. by-pass.

The circuit of Fig. 5 can be made up in a small metal box which will also house the battery, and by this means the pre-amplifier, which may be mounted several hundred yards from the set, can be energised and controlled at a convenient point—usually adjacent to the receiver.

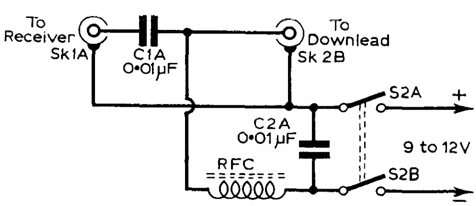


Fig. 5. The circuit of the local control-unit and power source.

Systems

Fig. 6a shows how two self-contained pre-amplifiers may be used in order to amplify the signals in Bands I and III. Here it is assumed that separate downleads are employed and that these are combined to a common set socket through a diplexer.

If a common downlead is used, as from a combined aerial, for example, then two diplexers are required to separate the signals for individual amplification, as shown in Fig. 6b.

Fig. 6c shows how two pre-amplifiers may be energised remotely through a common distribution cable or extended downlead. There must, however,

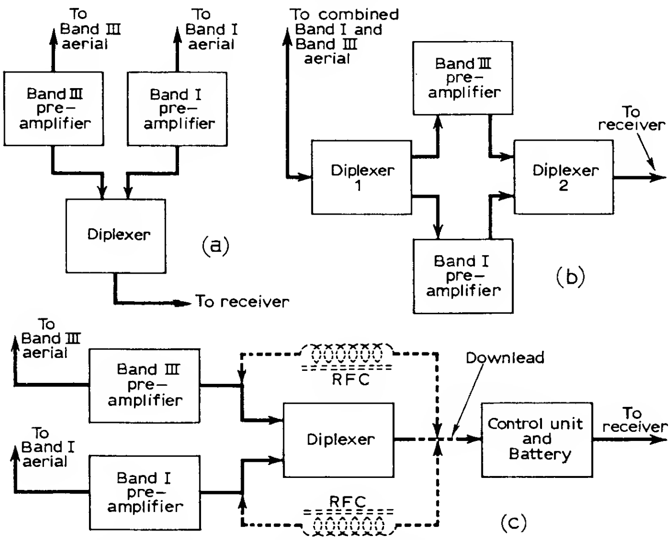


Fig. 6. Applications of the pre-amp: (a) two self-contained pre-amplifiers used for two-band amplification where separate downleads are used for BBC-1 and ITV; (b) two pre-amplifiers used where there is a common downlead for BBC-1 and ITV signals; (c) using remote powering to enable the pre-amplifiers to be used at the aerial-end of the downlead.

be continuity from the common socket of the remote diplexer to the Band I and Band III outputs (or inputs, when the diplexer is used round the other way). If continuity does not exist between the inner conductor terminals of the diplexer employed, small r.f. chokes must be connected between the inner conductor of the downlead and the Bands I and III inner conductor terminals of the diplexer as shown. The chokes provide a d.c. path for the powering current and have little effect on the operation of the diplexer.

If it is required to power a second amplifier at the set end of the downlead from the control unit and battery (Fig. 5), as shown in Fig. 7a, slight alteration to the input of the second amplifier is required, as shown in Fig. 7b. This allows the voltage on the line to be fed into the second amplifier via

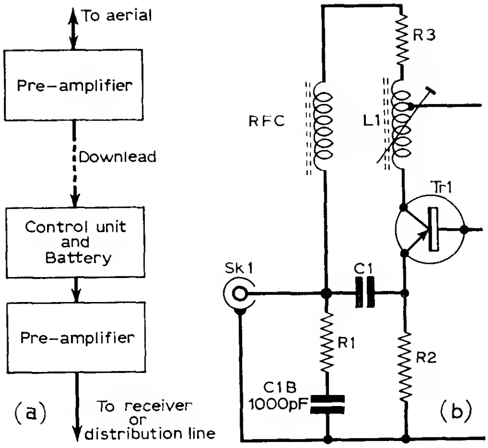


Fig. 7. (a), Block diagram of the arrangement required where two pre-amps are needed, one at each end of the downlead—the control unit supplies power to both pre-amps; (b), in order to use the arrangement shown in (a), the input circuit of the second pre-amp should be altered as shown here.

its signal input socket. The extra components are the r.f. choke (RFC) and the isolating capacitor C1B. For this application, it is also necessary to remove C1A from the circuit of Fig. 5. This, then, allows battery power to be fed from both sockets. Power-feed filters must only be used where line powering is required, because the chokes, although of fairly high impedance at signal frequencies, nevertheless by-pass some of the signal, particularly at Band III frequencies.

In Fig. 8a is shown how sets may be fed from the distribution line. Two resistors are used, and R2 usually has a value of 100Ω while the value of R1 determines the amount of signal fed to the set—680Ω gives about 20dB of attenuation, 330Ω 15dB and 180Ω 10dB. The value selected, of course, depends on the strength of the signal at the point on the cable where the tap is made.

The end of a downlead or distribution line may be terminated as shown at

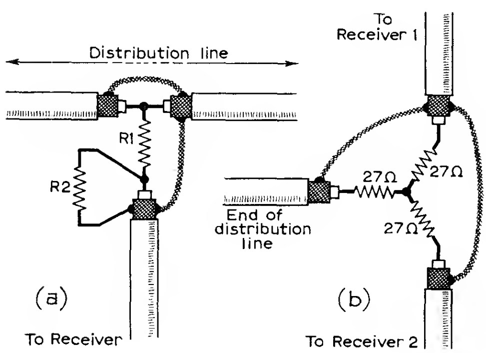


Fig. 8. (a), Feeding a series of receivers from the amplified signal in the distribution cable; (b), terminating the end of the distribution cable to feed two receivers.

Fig. 8b to feed two receivers. Each receiver in this case will receive approximately half the signal which is present at the end of the line.

It should be noted that loading the cable as at Fig. 8a adds to the overall losses in the cable. This loading loss, which averages 0·5dB per tap with the values for R1 given above, must be taken into account when working out the required amplifier gains.

List of Components

Resistors:
- R1 100Ω
- R2 1·2kΩ
- R3 470Ω
- R4 10kΩ
- R5 2·2kΩ
- R6 1·2kΩ
- R7 470Ω
- R8 10kΩ
- R9 2·2kΩ
- R10 100Ω
- (¼W 10%)

Capacitors:
- C1 500pF
- C2 1000pF
- C3 68pF
- C4 22pF
- C5 1000pF
- C6 68pF
- C7 500pF
- C8 0·01μF

These capacitors can be either midget or disc ceramics

Two coil formers with type "A" cores (Radiospares), 6ft of 26 s.w.g. enamelled copper wire and 3ft of 20 s.w.g. enamelled copper wire

3 five-way tag strips, 2 coaxial sockets, tobacco tin, 2 soldering tags, nuts, bolts, etc.

Tr1, Tr2 AF102 (Mullard transistors) or equivalent (see text)

S1 Round-dolly toggle s.p.s.t. switch

DETAILS OF COILS

Channel	Turns	Wire	Spacing
1	12	26 s.w.g.	close-wound
2	11	26 s.w.g.	close-wound
3	10	26 s.w.g.	close-wound
4	9	26 s.w.g.	close-wound
5	8	26 s.w.g.	close-wound
7	3·5	20 s.w.g.	⅛in between turns
8	3·5	20 s.w.g.	
9	3	20 s.w.g.	
10	2·75	20 s.w.g.	
11	2·5	20 s.w.g.	
12	2	20 s.w.g.	

List of Components
Control Unit (for Fig. 3)

C1A, C2A 0·01μF
S2, Round-dolly toggle d.p.s.t. switch

One 1A TV suppressor choke (RFC)
2 coaxial sockets, suitable box, wire, etc.

Chapter Four

A WIDE-BAND V.H.F. PRE-AMP

FOR some time now, there has been a need for a medium-gain wideband v.h.f. set-side amplifier, covering Bands I, II and III, from about 40Mc/s up to about 220Mc/s with a gain of the order of 12dB (16 times power amplification and four times voltage amplification). Such an amplifier can be useful for connecting between the common coaxial downlead and the television receiver's aerial socket, especially in those areas of weak and mediocre signal field and where the common downlead carries BBC and ITV television as well as Band II f.m. signals, as shown in Fig. 1.

An amplifier of this kind would also have a very useful application for supplying the whole range of signals to a number of receivers from a single set of aerials, as shown in Fig. 2. Nowadays, of course, quite a number of television sets feature f.m.-radio facilities, so an amplifier with gain over almost the entire v.h.f. spectrum is ideal for boosting signals for applying to this type of set.

There is no reason why, on the other hand, it could not be used for boosting the signals of just one Band I or Band III television channel or even the f.m. signals alone. It thus has a versatile application.

The amplifier to be described is of this type. Moreover it lends itself to ease of cascading for added gain and for line-powering in extended aerial systems.

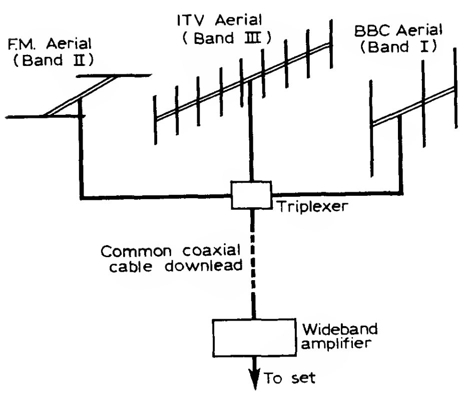


Fig. 1. The amplifier can be connected to a common downlead to amplify all signals simultaneously.

Fig. 2 (right). The amplifier may also be employed to feed signals in all channels to six receivers, as shown.

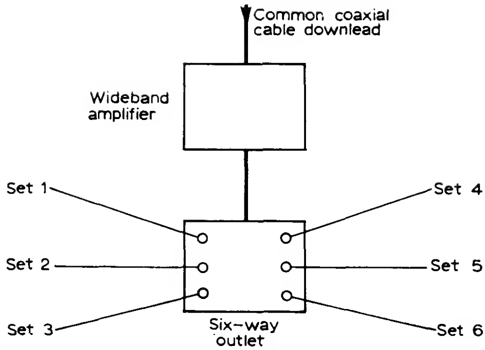
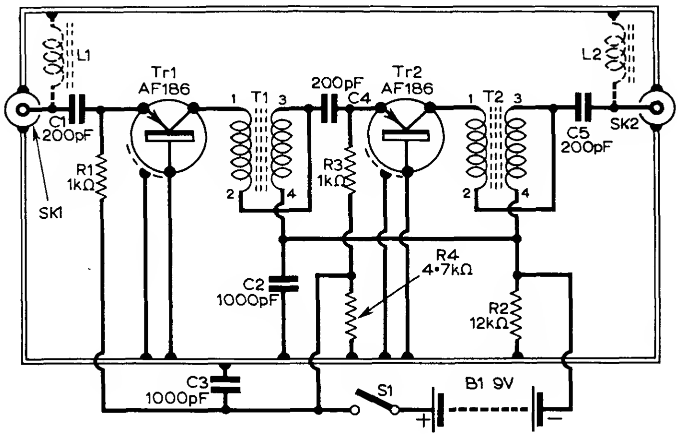


Fig. 3 (below). The circuit diagram of the amplifier (its operation is described in full in the text).



Circuit Description

The circuit of the amplifier is shown in Fig. 3. This is well worth consideration as it differs substantially from most ordinary v.h.f. transistor amplifiers. Two transistors are used, Tr1 and Tr2, each connected in the common-base mode. That means that the base is common to both the input and output signals. The signals are thus applied between the emitter and base and taken from the collector and base.

Tr1 is coupled to Tr2 via the special wideband transformer T1, while the signals are taken from Tr2 via a similar transformer T2. Neither of these transformers is tuneable. This makes the amplifier easy to construct and operate as alignment problems are totally avoided. It is absolutely non-critical.

Windings

The transformers can in effect be considered as extensions of an ordinary transmission line. The windings of this kind of transformer are arranged so

that the interwinding capacitance represents a component of the characteristic impedance of the transmission line. The inductive component is produced by winding the sections on a ferrite toroid in bifilar fashion.

"Bifilar" simply means that the two windings are wound together in parallel, the ends of the windings then being sorted out afterwards to provide the correct phasing and so forth. This method of construction avoids resonances which, in conventional transformers, can limit the overall bandwidth. The core permeability in conjunction with the number of turns on the windings governs the low-frequency response. This means that fewer turns are required for a given low-frequency response with a high permeability core material compared with a transformer made with a core of lower permeability.

The permeability of ferrites drops at v.h.f. but the efficiency of the transformers is maintained at the top end of the v.h.f. spectrum by the normal effect of the increase in frequency holding up the capacitive reactance between the windings.

T1 and T2

T1 and T2 are of identical design, the exact mode of construction being revealed in Fig. 4. Here it will be seen that one winding starts at 1 and finishes at 2, while the other winding starts at 3 and finishes at 4. The two windings can be identified as one is drawn in thick line and the other in thin line.

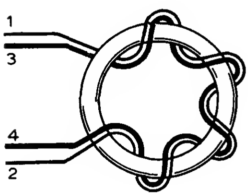


Fig. 4. Winding details of the ferrite transformers T1 and T2. The numbers on the windings correspond to those on the transformers in the circuit.

The two windings are wound in parallel with each other to give the bifilar attribute. There are four complete turns on the ferrite toroid. The toroid itself is nothing more than a simple ferrite bead costing about 2d. There are various types and sizes of such bead and the type employed is the kind designed for threading on to a conductor for increasing its inductance. This artifice is often adopted in v.h.f. equipment for "blocking" v.h.f. signals from power supply circuits and for similar applications.

This sort of bead is about $\frac{3}{16}$in in length by about $\frac{5}{32}$in in diameter with a hole through the centre of about $\frac{1}{16}$in diameter. It is very small and there is not a great deal of room for the wire. Nevertheless the windings can be accommodated without difficulty and the small size is technically desirable as it keeps proximity-effect losses at a low level.

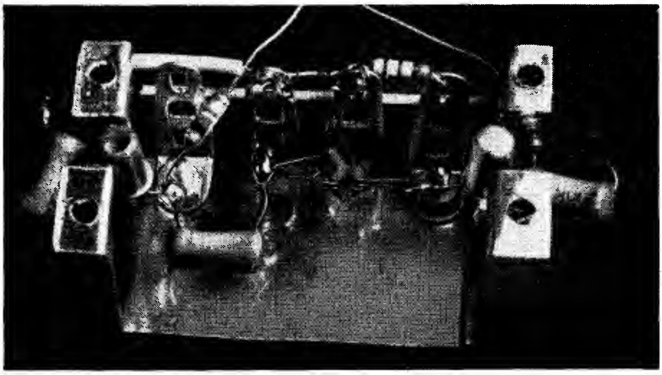
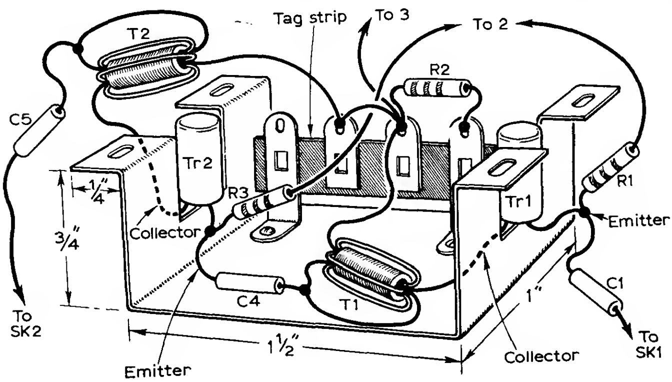


Fig. 5. Details of the sub-assembly. Note the ferrite transistor T1 connected between the transistors. The diagram below shows details of the wiring.



Wire

The wire for the windings is enamelled-covered 36 s.w.g. for winding 1–2 and 30 s.w.g. for winding 3–4. The two sizes of wire facilitate winding identification and ensure that the four double windings can be accommodated on the small size bead. Larger diameter wire should not be used. A finished transformer (T1) is clearly visible in the sub-assembly picture in Fig. 5.

The winding numbers in Fig. 4 correspond to the numbers against the transformer windings on the circuit in Fig. 3. It will be seen that on each transformer, ends 2 and 3 are connected together. This joint should be made as close as possible to the core of the transformer, connection then being taken to the circuit from this transformer winding joint. On no account should the joint be actually at the circuit connection as this would put two lengths of wire in series with the windings. The high-frequency response would then be impaired.

The way that the windings are connected endows the transformer with a 4-to-1 ratio impedance step-down. This means that the impedance at the

collector is stepped down four times or, conversely, that the impedance at the output of the transformer (i.e. as at the emitter circuit of Tr2 or across socket Sk2) is "seen" by the collector be be four times as high. It is by virtue of this impedance transformation that amplification is provided.

It is instructive to see just how this happens. When a signal is applied to the emitter circuit, as in the common-base mode, the signal current flowing in the collector circuit is approximately equal to the signal current as applied to the emitter. The input signal power P_{in} is thus equal to

$$P_{in} = i^2.Z_{in}$$

where i is the signal current and Z_{in} the transistor input impedance in the common-base mode.

Similarly the output signal power P_{out} is equal to

$$P_{out} = i^2.Z_c$$

where i is the collector signal current and Z_c the impedance as "seen" by the collector.

The power gain P_g is thus equal to P_{out}/P_{in} or

$(i^2.Z_c)/(i^2 \cdot Z_{in})$, which is equal to Z_c/Z_{in}

Thus if Z_c is equal to Z_{in} the power gain is unity. However, if Z_c is four times Z_{in}, as arranged by the 4-to-1 impedance transformer, then the power gain is four times or 6dB, while the voltage gain is two times. Two cascaded stages will give 2×6dB (12dB) or a power gain of 16 times (4×4) and a voltage gain of four times (2×2).

Transistors

This is the basis of design of the amplifier in Fig. 3. The input and output impedances are taken to be of the order of 50/70Ω. A number of transistors were experimented with and the Mullard AF186 was found to be highly suitable. In the common-base mode and adjusted for an emitter current of the order of 2·5mA (at 9V) the input impedance is around 50Ω. If the collector impedance is 200Ω, as reflected by the wideband transformer, the gain of each stage is as calculated above.

In practice the source impedance (i.e. the impedance "seen" by the transistor input circuit) and the output impedance (i.e. the impedance present across the output of the wideband transformer) are more like 70Ω, assuming ideal matching to the coaxial cables and freedom from standing waves. Thus the gain is slightly different from that calculated above, depending upon the extent of mismatch.

Matching Problems

Clearly the gain will be greater if the impedance to which the output of the amplifier is connected is greater than the source resistance at the input and, conversely, the gain will be smaller if the input source impedance is greater than the impedance "seen" by the output terminals of the amplifier.

A great deal of attention was given to the question of matching and attempts were made in an endeavour to present the emitters of the transistors with a source impedance of 50Ω from a 70Ω coaxial cable. This sort of matching is not impossible but it is difficult over a very wide range of frequencies.

Many tests made on this and a host of different transistorised v.h.f. amplifiers have revealed the futility of designing for a 100% impedance match when the equipment is to be used with an average television set and aerial system.

Although v.h.f./f.m. and television aerials are designed to load the feeder cable to match its characteristic impedance, say 70Ω, in practice the dipole impedance is rarely exactly equal to the characteristic impedance of the downlead, particularly in the case of wideband and composite aerial systems and when separate aerials are connected to a common downlead through certain types of diplexer and triplexer! This is not a criticism of aerial design but it reveals certain compromises that cannot be avoided in "domestic electronics".

While an aerial manufacturer may go to great pains with matching stubs and similar devices to ensure that, erected under ideal conditions, his aerial shows its feeder a perfect match, the whole equation is in practice unbalanced due to the need these days to erect aerials very close to neighbouring aerials on a shared chimney stack. The proximity of other large bits of metal does a great deal to destroy the matching that the manufacturer spent a lot of money and time in making optimum.

Cables

Other mis-matches occur when a flylead of coaxial cable slightly different from that of the main downlead is connected between the television or f.m. radio receiver and an outlet socket mounted on the window-sill, for instance. Bad mismatches can also occur in diplexers and triplexers as already intimated.

Moreover, while the aerial input at the receiver may be nominally valued at 70Ω or so, this impedance is not consistent over all channels (or rarely so). Indeed tests of admittance performed by the writer some years back showed changes of impedance ranging from about 30Ω up to 100Ω over the channels. This gives trouble in coaxial-relay and shared-aerial systems.

Nevertheless for optimum noise performance the best impedance match possible is necessary. If, for instance, it is required to use the amplifier essentially for lifting a very weak signal, say on Channel 9, an external matching arrangement could be connected between the aerial downlead and the input of the amplifier. If there is a slight mismatch on the Channel 9 aerial system then the simplest way of matching is by carefully adjusting the length of the downlead until the impedance that it shows to the amplifier is in fact 50Ω.

The downlead should be cut about an inch or less at a time and tried in the amplifier at each cut, aiming for the best noise performance (i.e. least

grain or "snow" on the picture). This technique applies to any transistorised amplifier.

The impedance variation at the receiver aerial terminal is less important. It is best to "swamp" it as much as possible by connecting the amplifier to the receiver through a piece of coaxial feeder of good quality not less than 9ft in length. This again applies to all transistorised amplifiers which are output-impedance sensitive—and most of them are.

This technique should also be adopted if there is any tendency for the amplifier to develop instability. In nine cases out of ten an amplifier-to-set coaxial lead not less than 9ft long will solve the problem.

Quarter-wave Transformer

An alternative method of matching an ordinary downlead to 50Ω is by the use of a quarter-wave coaxial transformer. This simply consists of the appropriate length of cable inserted between the downlead and the input to the amplifier. The characteristic impedance Z_t required by the coaxial transformer is equal to

$$Z_t = \sqrt{(Z_1 \times Z_2)}$$

where Z_1 is the characteristic impedance of the downlead and Z_2 the input impedance of the amplifier. Cable with a characteristic impedance of about 60Ω would give a reasonable match at the required "low-noise" frequency.

Now let us investigate the construction of the experimental amplifier. The prototype was built into an ordinary 2oz tobacco tin. The body of the amplifier is in the form of a sub-assembly, and the dimensions and make-up of this sub-assembly are shown in Fig. 5.

It consists of a piece of ordinary, thinnish tinplate (cut from a second tobacco tin or cocoa tin) cut to 1in in width and bent to have ⅜in sides, a 1½in top and ¼in top and ¼in flanges each side for fixing to the tobacco tin. Before the tinplate is bent, however, slots are cut in each side to accommodate the transistors as Fig. 5 shows. (Thin copper sheet may be employed instead of tin plate, and may give better results.)

The top of the sub-assembly is drilled so that a four-way tagstrip can be mounted inside. The two outside tags are "earths", the inner two being isolated. The tagstrip should be cut to have a length just equal to the inside of the bent tin so as to give it rigidity. The flanges are drilled to take 6B.A. bolts.

Mounting the Transistors

After the sub-assembly has been drilled and the tagstrip fixed, the transistors should be mounted, wire-ends down, as shown in Fig. 6. This diagram identifies the lead-out wires (looking at the base of the transistor). Note that both the shield and the base lead-out wires are soldered to the metal ends of the sub-assembly, giving the emitter lead-out on one side of the metal and the collector lead-out on the other. In this way almost perfect screening between input and output is assured.

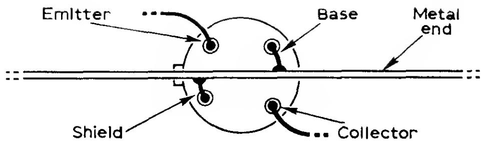


Fig. 6. The connections of the transistors showing how they are mounted.

Fig. 5 shows that the collector lead-out of Tr2 is outside the sub-assembly, while that of Tr1 is inside. Thus inside there is the emitter lead-out of Tr2 and the collector lead-out of Tr1. This facilitates the inter-transistor coupling via T1 and C4.

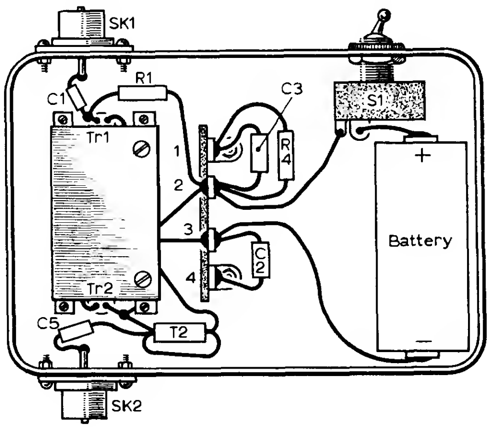
When the sub-assembly is wired to the extent shown in Fig. 5 the tobacco tin can be prepared to accept it. Fig. 7 gives the idea. Firstly the tin is drilled to take the sub-assembly, the input and output coaxial sockets, Sk1 and Sk2 respectively, tagstrip 2 and the on/off toggle switch. Tagstrip 2 is similar to that employed inside the sub-assembly, having two outer "earths" and two inner isolated tags. Fig. 7 also shows the point-to-point wiring of the sub-assembly to the various tin-located components and tags.

Finally clips suitable for a PP4 (or equivalent) 9V battery are soldered to the wires from C2 and S1 and to prevent the battery terminals from shorting against the metal tin, a length of thin rubber or cork should be positioned round the battery.

Testing

Before connecting the battery and switching on, the circuit should be thoroughly checked for correct continuity and to ensure that there are no short-circuits as these could be expensive! When completely satisfied in this respect the battery can be connected and the unit switched on. The total current consumption at 9V will be of the order of 5mA. Currents differing

Fig. 7. The wiring of the complete unit.



substantially from this would indicate a fault condition, in which case the unit should be switched off, the battery removed and the wiring rechecked. Note that the metal tin is common to the bases of the transistors and *not* to the battery. Neither battery terminal must be allowed to come into contact with the tin.

If the current is normal, the unit can be connected to a television set in series with the coaxial downlead, when a dramatic increase in signal strength will be noted on all programmes. The extra signal strength on f.m. will tend to reduce the amount of interference due to electrical effects such as car ignition systems, vacuum cleaners and so forth, since the receiver will now be pushed harder into amplitude limiting.

Transient Protection

It is rather important that any a.c./d.c. type receiver which is used with the amplifier be connected to the mains supply so that its chassis is at mains neutral. This is the correct way of connecting a set anyway. If the set is connected so that mains "live" is connected to the chassis there is the possibility that transient currents will flow through the transistor junctions each time the set is switched on and off.

These may well have a magnitude sufficient to destroy the transistors, even though the capacitor isolation in the set is well up to standard. This is because the transients are of an electrostatic nature, causing the charging and discharging of the input and output capacitors C1 and C5 via the isolating capacitors of the set.

The effect is considerably aggravated if the aerial happens to be earthed. One could then almost be sure that the transistors would fail either when the amplifier is connected or disconnected from the set or aerial or when the set itself is switched on and off.

One way of making sure that some protection is afforded should the set be wrongly connected to the mains supply is by connecting a small dust-iron core v.h.f. choke across the input socket and a similar choke across the output socket as indicated in Fig. 3 by the dotted inductors L1 and L2. Ordinary 1A TV suppressor chokes are ideal for this purpose.

They are virtually open-circuit at v.h.f. and an almost (but not quite) short-circuit to transients. They also serve to discharge any static build-up in the aerial system during thundery weather, which represents another source of voltage transients.

Cross-Modulation Performance

The emitter junction of any transistor amplifier is biased for forward current. A potential-divider, R2/R4, in this circuit performs this function for both transistors. The emitter junction, which is really a semiconductor diode, is very non-linear if insufficiently biased. Non-linearity at the input of a wideband amplifier which is called upon to handle a multiplicity of signals

is highly undesirable, for it could result in bad cross-modulation. This is when one signal modulates another.

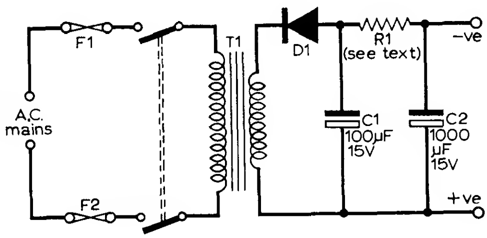
In practice this could cause bad pattern effects and inter-channel break-through, including sound-on-vision and vision-on-sound. The amplifier has been tested for cross-modulation and it has been discovered that with the transistors mentioned an output of at least 30mV can be obtained on six TV carriers and three f.m. carriers with cross-modulation still below 46dB (200 times down on the carrier signal voltage). It is thus invisible! The amplifier has a good application therefore for communal aerial and shared aerial systems.

Mains Power Unit

This is neither a difficult nor costly problem since only a small, very low-power supply unit is needed. A circuit of a unit of this kind is shown in Fig. 8. This can be built into any small box, even a partnering 2oz tobacco tin, quite easily. The biggest component is the mains transformer. This must have separate primary and secondary windings, and the primary winding must be suitable for the local mains supply voltage. The secondary winding produces a low a.c. voltage.

The amplifier will operate quite well over a range from 6V to 12V, but too low a voltage will limit the signal output level for a given ratio of cross-modulation. A transformer with a secondary output of between 6V and 9V

Fig. 8. The circuit of a mains-operated power unit suitable for the amplifier.



is suitable, for the d.c. output voltage will be a little above the a.c. voltage across the secondary winding. This is because of the very small current taken by the amplifier which allows the reservoir (C1) and smoothing (C2) capacitors to charge to almost the peak value of the a.c. across the secondary.

The voltage drop across the filter resistor R1 reduces the output voltage a little, depending upon the value chosen for the resistor, but, even so, the net output voltage is generally a little above the applied a.c. voltage. R1 can be chosen to have a value to provide an output of 9V with the pre-amp connected, or not less than 6V, as determined by the a.c. voltage across the transformer secondary.

In the prototype, a transformer giving about 8V a.c. and a 1kΩ filter resistor were used. With the electrolytic values as shown this gives good smoothing, as would be expected. A small heater transformer can generally

be over-run a little on the primary to give 7V or 8V across the secondary. Alternatively, one of the miniature transistor-equipment mains transformers can be employed. These are available with almost any required secondary voltage.

Diode

A small germanium diode is suitable for the main rectifier, shown as D1 in the circuit. The author has used an OA81 diode in this position quite successfully. This is capable of giving an average forward current of 17mA at 75V at a temperature as high as 75°C.

The primary is best connected to the mains supply through a pair of fuses, F1 and F2, valued at about 250mA, and an on/off switch (d.p.s.t.), S1.

If this sort of mains power supply is to be used with the amplifier there is no need for the amplifier switch proper. This can be deleted and the amplifier switched by the mains toggle, S1. A short length of well insulated twin conductor should be used to couple the power pack to the amplifier, the cable being protected from the tin cases by rubber grommets.

It is possible to build the power supply unit actually in the tin box of the amplifier if required, but this calls for one of the miniature mains transformers, mentioned above.

Line Powering

There are applications which demand the amplifier to be employed at a distance from the receiver and probably out of easy reach of a mains supply. In these cases it is best to power the amplifier over the coaxial cable carrying the signal. This is possible by the use of a mains filter at the power unit and by slight rearrangements at the amplifier.

A simple filter is shown in Fig. 9. This can be built into the case of the

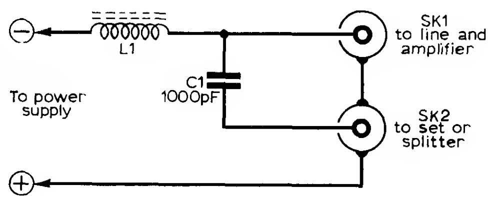


Fig. 9. A circuit of a power filter which may be incorporated in the power unit for powering the amplifier via the coaxial cable.

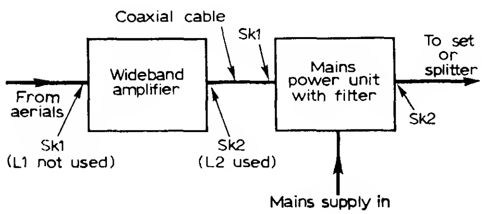
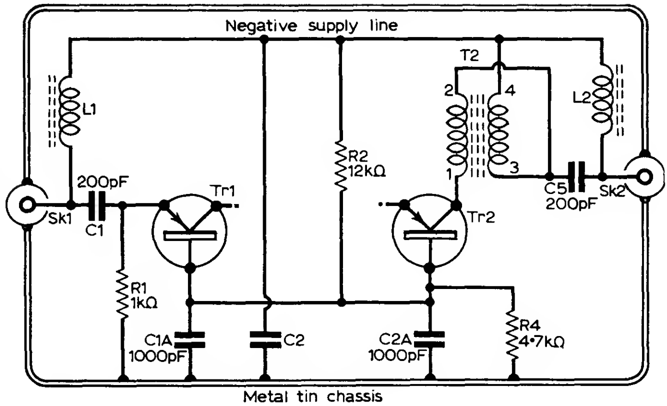


Fig. 10. When line-powering is employed, the amplifier is generally best arranged to be next to the aerials.

Fig. 11. The arrangement of the amplifier for line-powering. The capacitors C1A and C2A are extra components.



mains power unit. All it consists of is a filter choke L1 to pass the negative d.c. supply to the "line and amplifier" socket and a capacitor C1 to pass the signal from that socket to a socket which accommodates the coaxial lead to the set or splitter if the amplifier is used to supply a number of receivers from a common aerial (see later).

The method of connection is shown in Fig. 10. Here it will be seen that the d.c. supply from the mains power unit is fed via the filter from Sk1 (Fig. 9) to Sk2 of the amplifier. If we look at the circuit in Fig. 3 we see that as the amplifier stands, the d.c. would not get into the amplifier. Without the transient protection choke L2, the d.c. would be isolated by C5. With the choke L2 in circuit, then this would short-circuit the d.c. supply. On no account, therefore, must line power be injected into the basic battery-powered amplifier.

Amplifier Modification

To accept the line power, the amplifier must be altered a little. How this is done is revealed in Fig. 11. Here it will be seen the bases of the transistors are no longer connected to the chassis. Instead they are connected via extra capacitors, C1A and C2A, to the metal box. These capacitors should be mounted as close as possible to the base lead-outs and "earthed" to the metal sides of the sub-assembly. Ceramic capacitors should be used.

The two bases are connected together and they are biased by R2 and R4 as in the original model, but R4 is now connected to "chassis" and R2 to the *negative* line. This calls for a slight rearrangement of the wiring on the tag strips. C3 in the original model is no longer required, neither is switch S1.

This time, chokes L1 and L2 are essential, but instead of being connected to the metal chassis, they are connected from the centres of the coaxial sockets to the negative supply line. Looking again at Fig. 9 it will be seen that the positive of the supply connected to the outer conductor (braid) of

the coaxial cable feed to the amplifier. The choke L1 in this circuit connects the negative of the supply to the inner conductor. Power at the correct polarity is thus injected into the amplifier via the coaxial cable, socket Sk2 and L2.

The chokes represent an almost open-circuit to v.h.f. signals while passing the d.c. supply without appreciable resistance. The d.c. is blocked by C1 in Fig. 9 and by C5 in Fig. 11. These capacitors, however, pass the v.h.f. signals without undue loss. The chokes employed are ordinary 1A television suppression chokes designed for the suppression of electrical equipment against television interference. Suitable chokes are manufactured by Belling and Lee Limited, among others.

Extra Amplifiers

Choke L1 in the input circuit allows the power supply to be fed in via Sk1 if required, giving the arrangement shown in Fig. 12. Moreover, two amplifiers (or more if the power supply unit can handle the extra current without too great a voltage drop) can be powered from one line since the inner of the input socket is connected to the inner of the output socket, via the negative supply line and the two chokes. This permits the arrangement shown in Fig. 13, the two amplifiers being in cascade.

When an amplifier is line-powered it must be remembered that by the use of two filter chokes, one at the input and one at the output, the power supply is present at both ends of the amplifier. This can result in a short-circuit across the power supply under certain conditions. For instance, some aerials and diplexers or triplexers have d.c. continuity between the two conductors of the coaxial cable. Thus, with the two chokes connected in an arrangement as at Fig. 10 a short could occur on the power circuit. This is avoided by connecting the choke corresponding to the end of the amplifier that the power is to be injected and deleting the other one. In Fig. 10, therefore, only choke L2 would be used, while in Fig. 12 only choke L1 would be used. In Fig. 13, of course, both chokes would be connected.

An extended coaxial circuit with one, two or more amplifiers might be necessary to feed a multiplicity of v.h.f. signals over a distance of 200yd or more from a hill-sited aerial, for instance, to receivers in a valley. Without any amplification, the aerial signals would soon be weakened by the attenuation of the coaxial feeder cable. The attenuation is greater the higher the frequency of the signals. Thus, signals in Bands III and II are attenuated more over a given length of cable than signals in Band I.

If an amplifier is used at the receiver end of a long run of aerial feeder cable, the picture "noise" (i.e. grain and snow effects) will be greater than necessary due to the weakened signal being applied to the input of the amplifier. The strongest possible signal should always be fed to any amplifier to secure the best signal-to-noise ratio.

The use of an amplifier close to the aerial, where the signals are most powerful, is often made difficult if not impossible, by the lack of mains supply

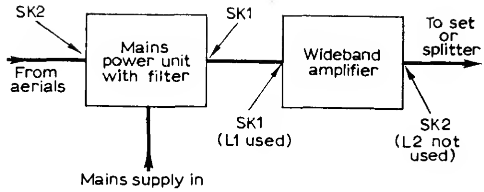


Fig. 12. It is possible to arrange the line-powered circuit so that the amplifier is next to the set or the splitter, as shown.

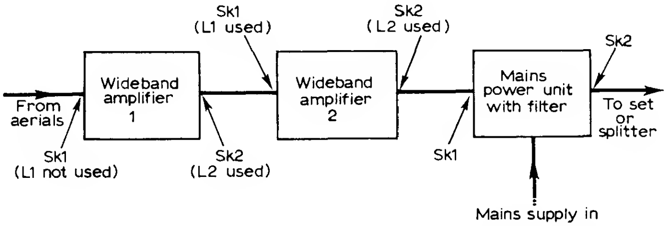


Fig. 13. Two (or more) amplifiers can be line-powered from a single supply unit as shown in this diagram.

in proximity to an isolated aerial site. The line-powering of amplifiers resolves these difficulties, of course.

Signal Splitting

To operate a number of receivers from a common cable carrying a multiplicity of signals, some means is necessary for splitting the common feed into the required number of set-feeds. It is not possible simply to parallel a number of feeders to a common cable as this would introduce mismatch effects and cause a great deal of trouble. Some means of matching each set-feed is essential whilst maintaining the correct matching on the common feeder.

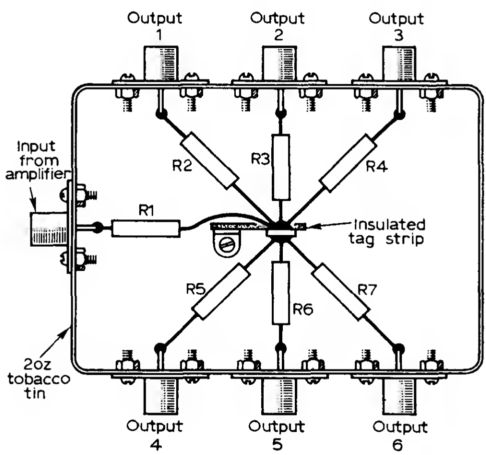


Fig. 14. Details of the construction of a six-way star network.

A simple way of getting a number of matched outlets is by means of a "star" network. A six-outlet star network is shown in Fig. 14. This simply consists of six coaxial sockets and six resistors for the outputs plus one coaxial socket and one resistor for the input. The device can be built easily into a 2oz tobacco tin to match the amplifier and power unit. In the centre of the tin is a tagstrip carrying one insulated tag at which all the resistors terminate.

Since the impedances of the input and output coaxial cables are equal, all the resistors must have the same value. The value depends upon the number of outlets and can be found from the simple expression $R = Zo(n-1)/(n+1)$, where Zo is the characteristic impedance of the cables (usually about 70Ω) and n is the number of outlets. For six outlets, R works out to be just a little under 36Ω. The nearest preferred value would probably be used in practice. This is 33Ω. Insulated resistors are most suitable for the application, and they should be of the non-inductive carbon type.

The insertion loss of such a device is equal to the number of outlets. Thus, the loss given by the six-outlet star will be six times. This means that if a signal of, say, 300μV is fed to it, only 50μV of signal will be present at each correctly matched output. Here, then, is revealed the reason why signal amplification is essential when it is required to feed a number of sets from a single aerial system.

Let us suppose that an aerial signal of 1,000μV is present on all required channels. Using one wideband amplifier would lift the strength of each signal to 4,000μV. If the amplifier output were then applied to a six-way splitter, each outlet would deliver a signal of about 660μV on all channels which would be adequate for most sets.

List of Components

Capacitors:
C1, C4 and C5 200pF ceramic
C2 and C3 1000pF ceramic

Resistors:
R1 and R3 1kΩ
R2 12kΩ
R4 4·7kΩ
All ¼W or ½W insulated

Transistors:
Tr1 and Tr2 Mullard AF186

Battery:
9V (PP4 or equivalent)

Miscellaneous:
2oz tobacco tin. Piece of sheet tin. Two coaxial sockets. S.P.S.T. toggle switch. Two ferrite beads. Lengths of 30 and 36 s.w.g. enamelled-covered copper wire. Two four-tag strips. Battery clips (for PP4 battery). 6B.A. nuts and bolts. Connecting wire and battery leads

Chapter Five

A U.H.F. PRE-AMP

THE start of a u.h.f. television network in Britain brought some interesting problems, but one of the most difficult to solve was the higher noise level of the received picture in many areas compared with the pictures from the v.h.f. television signals. However, the introduction of new types of transistor made possible very simple and efficient amplifiers for u.h.f. signals. The Mullard AF186 *pnp* alloy-diffused transistor is one and is employed in this pre-amplifier. The transistor was found to work very well indeed in a simple, easily constructed, $\lambda/4$ tuned cavity.

A common-base circuit is used (Fig. 1) and the input is untuned because the emitter is a reasonable match to the aerial cable and better matching is hardly worth-while. The base circuit is rearranged so that the collector circuit is all "d.c. earthy". This simplifies the amplifier. The output is tapped directly into the tuned cavity, no particular position. Moving the tap towards the transistors increases the stability and bandwidth but reduces the gain.

Quarter-wave tuning lines can be twin or coaxial and this one can be considered as coaxial with the outside part made of square section for ease of construction and with one long side omitted. The box is made from a piece of 21 s.w.g. copper approximately 3in $\times$ $4\frac{1}{4}$in. Squares are cut from each corner and the four sides are then folded up and the four corner seams soldered, making a box $2\frac{1}{4}$in $\times$ 1in (see Fig. 2). The inner part of the tuner consists of a piece of $1\frac{5}{16}$in $\times$ $\frac{1}{8}$in diameter copper tube or rod, one end of which is soldered to the centre of the end of the box and the other end is soldered to the "hot" end of the trimmer. This tuning capacitor is of the type consisting of a threaded ceramic tube into which a brass screw is inserted. The screw forms one plate of the capacitor and a tinned silver ring on the outside, the other. The silvered ring on the outside is the "hot" side of the trimmer. These are widely used on v.h.f. tuners and in this amplifier it also makes a convenient support for one end of the $\frac{1}{8}$in copper rod.

The inner contact of the output socket is soldered to the copper rod.

The transistor is push-fitted into a suitable hole and the feed-through capacitors are soldered into suitable holes each side of the transistor. Leads should be kept short, particularly the base lead to the feed-through capacitor. Complete all soldering operations to the box well before fitting the transistor as the box becomes very hot.

A small choke consisting of five turns of enamelled wire wound on a $\frac{1}{4}$in former should be fitted across the input of the amplifier to prevent the application of large and possibly destructive pulse voltages to the transistor.

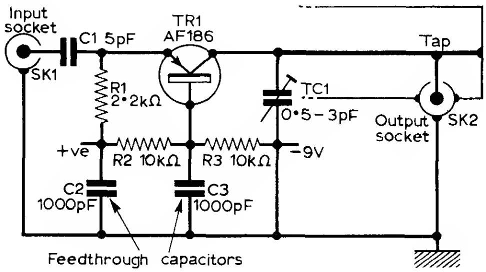
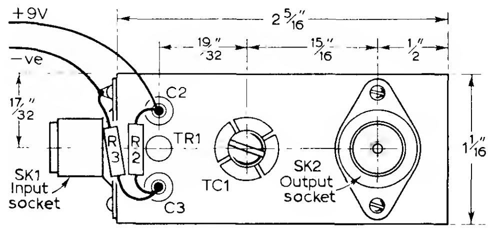
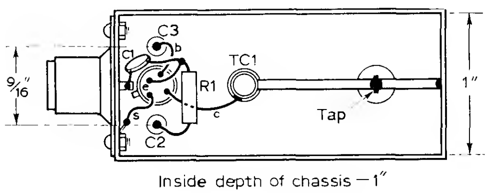


Fig. 1. The circuit of the u.h.f. pre-amplifier.

Fig. 2 (right and below). Top and underside views of the pre-amplifier.





For best results, the amplifier should be mounted close to the aerial, as amplified signal and amplifier "noise" are then attenuated by the cable, whereas, if the amplifier is at the set-end of the cable only the signal is attenuated. For low-loss cable runs, the difference is small.

The optimum operating current for this transistor is 2mA, and although the voltage supply is nominally 9V the amplifier works well even when the supply voltage is below 5V. The dimensions are shown in Fig. 2.

Laboratory measurements have shown a gain of better than 18dB, a noise factor of 7dB and a bandwidth in excess of 35Mc/s.

List of Components

Resistors:
R1 2·2kΩ
R2 10kΩ
R3 10kΩ
All 10% ½W

Capacitors:
C1 5pF ceramic
C2 1nF (1000pF) feed-through type
C3 1nF (1000pF) feed-through type
TC1 0·5pF–3pF trimmer

Transistor:
Tr1 AF186

Copper for chassis (see text); two coaxial sockets; etc.

Section 2

RECEIVERS

Chapter Six

THE "OLYMPIC II" TRANSISTORISED RECEIVER

THIS chapter will describe in full detail the specification, construction, adjustment and performance of a fully-transistorised television receiver for the home constructor. Naturally, a certain amount of experience with both transistor and television receivers has to be assumed, but although not suitable as an exercise for the beginner, no difficulty should be found by one who has built a few transistor circuits and knows something about TV from the practical angle. A certain amount of theory will of course be mentioned, if only to interest those who will later on carry out a conversion of the circuit; the need for this will be explained also.

The present design is for a straightforward 405-line Band I and Band III receiver. Band V has not been neglected during the design procedure, however, and the circuit is such that conversion to the 625-line standards can be done readily when circumstances make it desirable. (Specification page 120.)

General Description

Fig. 1, which shows the block diagram of the 405-line receiver, indicates that in essence the circuit follows conventional lines. The tuner unit comprises a radio-frequency amplifying stage consisting of a v.h.f. transistor operating in the common-base configuration, together with a similar transistor operating as a common-base oscillator. A further v.h.f. transistor acts as a frequency changer, again operating in the common-base mode, and the i.f. signals (38·15Mc/s sound, 34·65Mc/s vision) are taken from its collector circuit by means of a suitable transformer and passed to the i.f. amplifiers at the appropriate impedance level.

The sound i.f. amplifier consists of two slightly under-coupled transistor stages, somewhat over-neutralised in order to sharpen the selectivity and improve the sensitivity. This has been found to economise in one v.h.f. transistor at the expense of a little extra care in setting-up. The vision signal is completely eliminated in this i.f. amplifier; although the selectivity on the low-frequency side of the response curve is not so high as on the high-frequency side, no trouble has been found from adjacent-channel vision.

The detector is the usual germanium diode, and between this and the high gain audio amplifier is interposed a noise-limiting device. Because of the

low impedance levels which are inescapable with transistors in practical circuits, this is not so effective as in a similar valved circuit, but the difference in performance is not great and the effect of car ignition interference is not usually troublesome.

The audio amplifier consists of a pre-amplifier stage, followed by a 1W combination of driver and push-pull class AB output. The rating of 1W is a conservative one, and in practice $1\frac{1}{2}$W—2W is easily obtained; much depends here on the heat-sink arrangements for the output transistors.

The vision amplifier comprises three transistor stages at intermediate frequency, followed by a pair of transistors working as a conventional video amplifier. The use of a high-voltage video output stage enables plenty of modulation drive to be available for the cathode-ray tube.

The i.f. stages are preceded by a sound-trap giving only a small rejection notch at the sound intermediate frequency. The main sound-rejection is arranged to take place between the last two i.f. transistors, where an "infinite-rejection" circuit is interposed. The total rejection is nearly 50dB at the sound i.f., with sufficient rejection at the edge of the audio pass-band—over 42dB at $\pm$ 100kc/s about 38·15Mc/s. This eliminates sound-on-vision entirely, providing tuning is reasonably accurate, and the effect of the rejectors on the vision pass-band response is readily corrected by following a simple drill in adjusting the i.f. transformers.

The first two i.f. stages are coupled by a single-tuned circuit, while stages 2 and 3 utilise two tuned circuits "top-inductive-coupled" by means of the main sound rejector. The third stage feeds the detector via a single-tuned inductor.

The field time-base generator is based on a transistor blocking oscillator, synchronised direct by the field sync pulses derived from an interlace device. The output transistor is choke-coupled to the high-resistance field-scan coils, and linearity correction is incorporated.

The line generator is similarly a blocking oscillator, which provides the output to a driver stage.

The input to the driver is adjusted by means of a very stable tuned circuit, which varies the "on-time" of the oscillator, while the output is used to operate the line-output stage as a current switch. Direct coupling to the low-inductance scan coils ensures high efficiency, although in order to obtain the high e.h.t. a small portion of the yoke current is diverted to a "line-output transformer". This transformer allows lower voltages of the correct polarity to be obtained for the first and focus anodes of the cathode-ray tube and for the video transistor, and for control of brightness.

Because of hole-storage effects, it is not practicable to employ direct locking of the line oscillator, and this would in any case be inferior when conversion to 625-line standards is undertaken. Consequently a simple "fly-wheel" circuit is provided for line synchronisation, although more properly this should be thought of as a phase-correcting circuit.

The receiver uses a 14in electrostatically focused tube operating at 14kV and giving a maximum light output of 100ft lamberts—in fact this tube gives

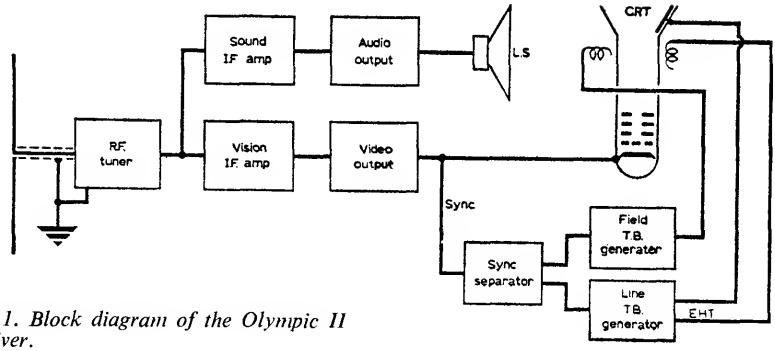


Fig. 1. Block diagram of the Olympic II receiver.

an exceptionally brilliant picture. This is a valuable feature, since the receiver is so light and robust that it can readily be transported in the boot of a car for use at picnics. Used in surroundings where ambient light is likely to be intense, a good bright picture is a great advantage.

Sensitivity is very high on both Band I and Band III; the performance in this sense is well up to the standard of commercial valved receivers. The noise level at maximum sensitivity is especially low; when properly set up, the noise is about half that found with a double triode r.f. amplifier in modern valved tuners. Provision is made for the reception of six channels at the moment.

The weight of the receiver depends to some extent on the actual construction adopted. The tube itself weighs about 9lb, and the weight of the components and wiring about 5lb or 6lb. If an 8½in tube had been used, it could have been a very small and light assembly, but most viewers would prefer a somewhat larger picture than this would give.

Power supplies have been thought out with some care. For most of its life, the set is likely to be used in the home as a mains receiver, and because of this a mains power supply has been designed for it. A 12V car accumulator will power it when used away from home, and if a small 12V accumulator is incorporated as an internal power supply, it will be independent—for an hour or two—of the power cuts with which "the economy" is threatened from time to time. Caravan dwellers without laid-on electricity will probably find that two car accumulators are the best means of providing power—one on charge while one is in use. The current taken is 800mA (0·8A), which represents a power consumption of about 10W.

Detailed Circuit Description

It will be noted that "unit construction" has been employed. This was not done as a deliberate policy, in spite of the advantages such assembly may have. In fact, each section of the receiver was conceived and designed separately, because new ground was being covered in many respects, and it just

turned out that way. It would have been perhaps better in some ways to have rebuilt the entire receiver into a specific form; but time would have been lost, and in addition flexibility would certainly have been impaired.

The prototype, as illustrated, therefore represents only one of a number of possible ways in which the sub-units may be assembled. It is true that certain layouts must be avoided—it would be very unwise, for example, to put the line-output stage farther from the cathode ray tube than the e.h.t. lead will reach without extension!—but these will be noted as the occasion arises. Apart from such considerations, the layout of the sub-assemblies is not critical, and as a bench hook-up, the most extraordinary positions have occurred without causing any trouble at all. There is no need at all to employ unit construction, and provided the screening is as specified the whole receiver can be mounted on a chassis in the conventional way. However, considerable advantages accrue from making a separate unit of the r.f. tuner, and as this is a relatively critical part of the receiver the constructor is recommended to follow the layout and assembly given here.

The r.f. tuner circuit is given in Fig. 2, and as mentioned earlier, the common-base configuration is used for all the stages. Although it is true that certain transistors now available would enable common-emitter working, the common-base arrangement has advantages in that neutralisation is not essential and cheaper transistors can be used. In any case, when the time arrives to convert to u.h.f., common-base operation will be essential.

The aerial is coupled directly into the emitter of the first (r.f.) transistor, Tr1, and this might be thought to represent a loss of match, liable to give rise to lack of sensitivity. In fact the loss is small if 75Ω cable and the appropriate aerial (plain dipole or the usual array having a 75Ω impedance) is used. The use of a π-coupling network to match the aerial to the transistor input certainly gives a small reduction of noise, but this would require switching for other channels and itself represents a source of some loss of signal, it was thought better to simplify the circuit, and any loss is not noticeable in practice.

The input circuit is not untuned, as might be thought at first sight. The aerial is itself the first tuned circuit, and this implies that its "Q" must be low enough to give negligible attenuation at the edge of the band to which it is tuned. Consequently a "throw-out" aerial consisting of flex will give poor resolution; the use of ¼in tubing has been found however to give excellent results, and larger diameters can be used with advantage. The length should be accurate for the channel to which the receiver is tuned, and if portable operation is likely a very suitable type is a telescopic dipole, such as may be purchased from a number of manufacturers. Nevertheless, at a distance of 22 miles from the Mendlesham ITV transmitter, a good picture was received on a small screwdriver plugged into the aerial socket, so it may be seen that one need not be too fussy after all. If the aerial is mismatched at the higher frequencies of Band III, however, it will be found that hand-capacitance effects may be noticeable, or that the tuning depends somewhat on the amount, and the position, of the coaxial lead. This clearly needs to be minimised, so it is worthwhile to take some trouble.

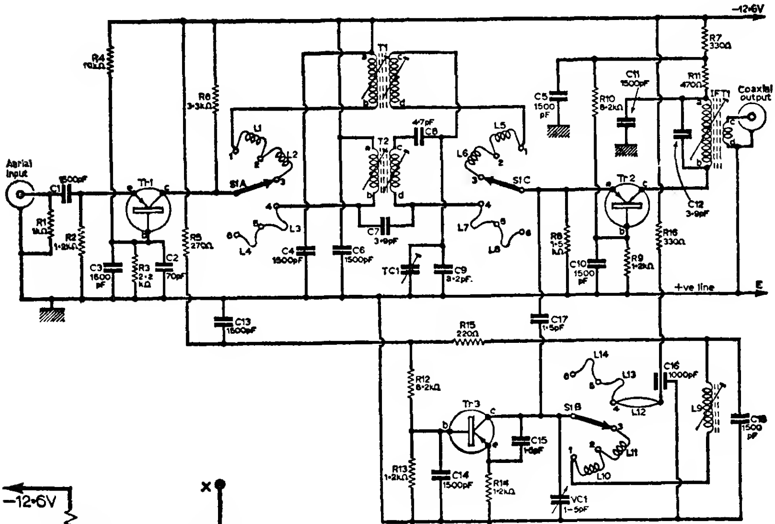


Fig. 2. The circuit of the r.f. tuner unit. Details of the transistors are given in the List of Components on page 121.

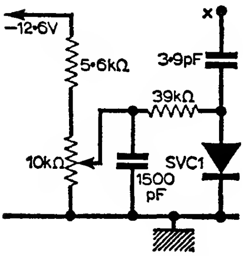


Fig. 3 (left). Electronic fine tuner circuitry.

Although a self-oscillating mixer could be used it was considered better' for several reasons, to use a separate oscillator. The oscillator used is about the simplest possible, and works satisfactorily at all the frequencies of interest, without changing the feedback capacitor (C15) between collector and emitter. The value of this capacitor is nominally 1·5pF, but is by no means critical. However, its value affects markedly the size of the Band III inductors, and it is best to use the specified value.

Tuned-transformer coupling is used between r.f. and frequency-changer stages, and the secondary of the inter-stage transformer is π-coupled to the f.c. emitter to effect impedance-matching. This matching is accomplished by the combination C9 and TC1, in conjunction with the input capacitance of Tr2, the f.c. transistor. Capacitor C9 is effective on Band I, whilst on Band III, TC1 permits optimum results to be achieved. The inter-stage coupling gives a band-pass effect which improves selectivity without impairing bandwidth, but the matching needed is not critical and best results are obtained by adjusting TC1 on a Band III channel to obtain minimum noise. This cannot be done readily unless the output from the tuner can be fed into a suitable i.f. amplifier, and may be left until the receiver is complete.

Fine tuning on any channel is accomplished by VC1, a 1–5pF variable capacitor in the oscillator circuit, but a modification may be incorporated,

which avoids the use of a variable capacitor and uses electronic tuning instead. This allows the use of a d.c. control, namely a potentiometer, and so enables remote tuning to be obtained if needed. The circuit is shown in Fig. 3: VC1 is removed, and point "X" in Fig. 3 is connected instead to the collector of the oscillator transistor Tr3 (pole contact S1b). The only extra connection is a lead from the −12·6V supply to the 5·6kΩ resistor. (The Mullard SVC1 is a variable-capacitance diode, available through one's local dealer.)

Channel-selection is by means of incremental inductances on both Band I and Band III. This requires some patience in setting-up, but a turret tuner is bulky and hard to come by. The alternative of separate switched coils is possible for Band I, but not so easy to arrange on Band III where the switch elements themselves form a major part of the tuning inductors. A choice of six channels is allowed by the nature of the switch specified: it is important to use this switch since it is small, electrically excellent and very stable in use.

The Band III "padding" inductors are merely short lengths of wire, and in the oscillator circuit, which operates at a higher frequency than the signal, L12 consists of two wires in parallel. The value of its inductance can be adjusted by varying the spacing between the "parallel" wires, and this is easily done by means of a Perspex strip such as is used for trimming iron-dust transformer slugs.

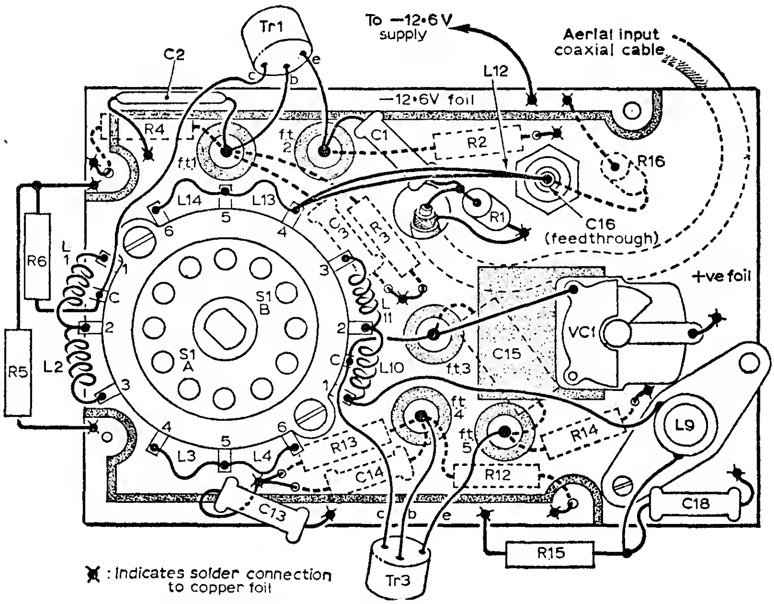


Fig. 4. Wiring on PC1 ('f.t.' indicates anchor tags).

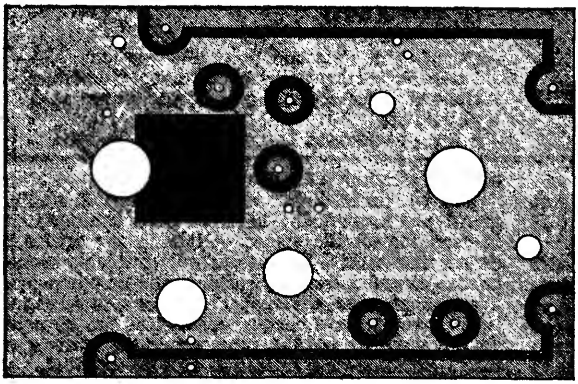


Fig. 5. Details of printed board PC1—the black *parts are etched away.*

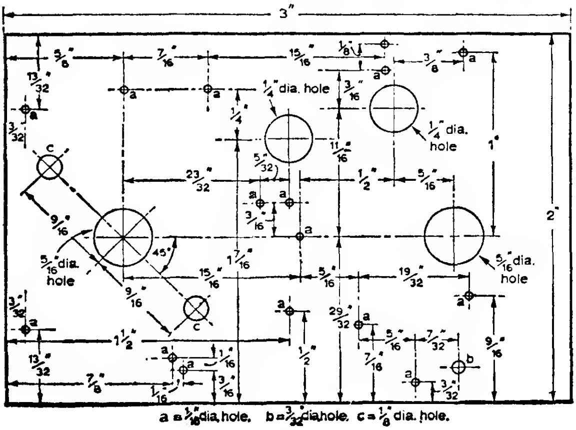


Fig. 6. Details of the drilling of PC1.

Constructional Details

Etched-circuit construction is used in the tuner, and the two small circuit boards are of epoxy glass-fibre laminate. This is electrically very stable and humidity changes have very little effect on its electrical characteristics. However, the cheaper E60 grade may be used if desired, as the DH74 is rather expensive. Bakelite Ltd. do not supply direct to the public, but application for the laminate may be made to Anglo-American Vulcanised Fibre Co. Ltd., Cayton Works, Bath Street, London, E.C.1.

The two printed circuit boards will be referred to as PC1 and PC2 respectively (Figs. 4–9). These may be marked out with resist direct, or a photographic method may be employed (see Chapter 11 for details of the methods).

Since interconnections between the boards are necessary, certain components are best mounted before putting the assembly together, and a few notes on this may be of use to the constructor. The layout is arranged so that each circuit board is mounted on the switch itself, closely adjacent to the

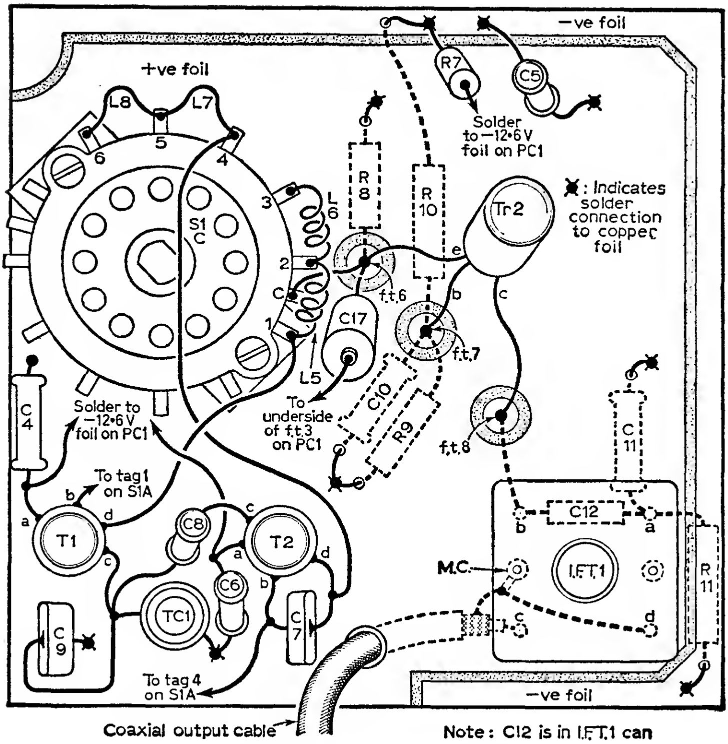


Fig. 7. Wiring details of PC2.

wafer concerned; the r.f. and oscillator stages are arranged on PC1 board which is farthest from the switch knob, while the frequency-changer goes on PC2 board which is nearest to the switch knob (see Fig. 10).

The minimum amount of copper is etched away—sufficient to form anchor tags for the transistors and a rail for the "battery minus" supply (−12·6V). The remainder then acts as screening between the stages. On the PC1 circuit board, a square area is also cleared of copper so that the minimum capacitance of the variable capacitor VC1 (oscillator fine tuner) is as small as practicable (see Fig. 5).

When the boards have been etched and cleaned up they can be mounted together in a vice and drilled for the switch holes (Figs. 6 and 9). The switch control hole is centred at $\frac{5}{8}$in from the top edge of the board and $1\frac{1}{8}$in from the right-hand side. The holes for the side struts are at 45° as shown and they are $1\frac{1}{8}$in apart. If both boards are drilled at the same time there will, later on, be no trouble about alignment. If the straightforward variable capacitance method of tuning the oscillator is used, the holes for this component should also be drilled in the same operation.

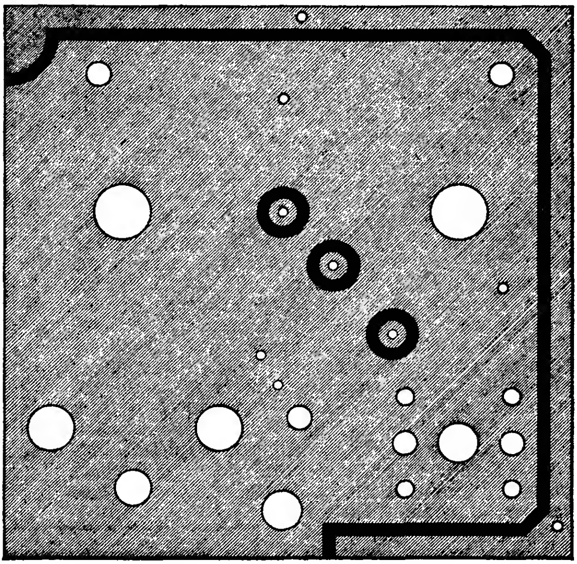


Fig. 8. Details of printed board PC2—the black *parts are etched away.*

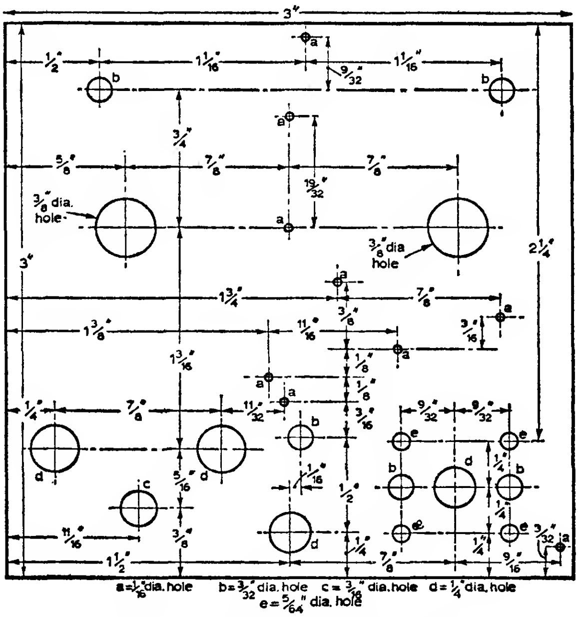


Fig. 9. Details of the drilling of PC2.

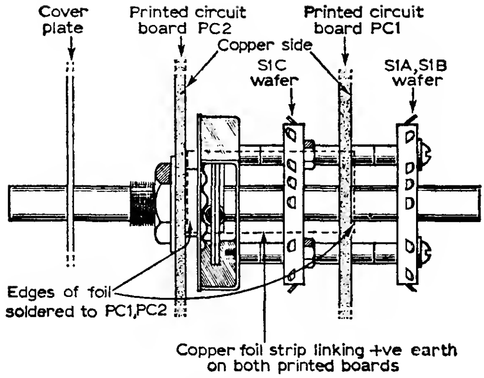


Fig. 10. Assembly details of the channel-selector switch.

The larger circuit board (PC2) is next drilled to take the inter-stage transformers T1 and T2. For these, the prototype receiver uses the 0·3in Bakelite tubular former from one of the "long" canned assemblies which are retailed widely for coil-winding. The base is cut off so as to give the maximum possible length, and it will be found a simple matter to fix these into the holes now drilled for the $\frac{1}{4}$in diameter portion of these formers.

The smaller circuit board (PC1) is next drilled to accept the Band I oscillator coil. This is a small polystyrene former of 0·3in diameter, and may be cut from a "short can" assembly in the same way. Finally, both boards are drilled with a $\frac{1}{16}$in drill to take the leads of the small components. In the layout of the small components (the resistors and capacitors affording the correct d.c. supplies to the transistors) it is sufficient to take them to the nearest convenient point, both for B– supply and for earthing. The separate board design ensures that the layout is not critical in this respect. There are, however, one or two critical points of layout, and these are detailed below.

The emitter of the first r.f. transistor (Tr1) is coupled directly into the aerial via a 1500pF capacitor C1, and any accumulation of static is prevented by affording a low-impedance d.c. path (R1). The coaxial cable is taken along the underside of PC1 circuit board, through a hole, and the coaxial outer is soldered to the board close to this hole. The "inner" joins direct to C1 and R1; anchor tags are not provided for these components as this might impair the screening. Besides decoupling the base of Tr1 with the 1500pF capacitor C3, a further 70pF capacitor is soldered to the transistor base lead as close to the transistor case as possible. Only enough space to get in the point of sharp-nosed pliers—to use as a heat shunt—should be allowed; the soldering must be done quickly with a very hot iron, and the joint should be cooled with a wet finger forthwith. The earthy end of C2 is then soldered to the board at the nearest convenient point. Keep all end wires of components as short as possible; e.g. C3 should be mounted as close as possible to the anchor tag ft1.

Only two anchor tags are provided for Tr1. The collector lead is taken

direct to the pole contact of S1a. This minimises lead inductances and stray capacitances. It may be found an unnecessary precaution, but is advised.

The oscillator transistor (Tr3) collector lead is similarly taken direct to the pole contact of S1b but an anchor tag (ft3) is provided. This is not for the collector of Tr3, but for the variable capacitor VC1, and the capacitor C17 which couples the oscillator to Tr2, the frequency-changer. If electronic tuning is used, the 3·9pF series capacitor (see Fig. 3) is taken direct to the pole contact of S1b also.

The transformers T1 and T2 are so wound that the primary winding is just above PC1 circuit board and the secondary winding just below. This enables the B− contacts to be made at the edge of PC1 board, with the shortest leads to the switches.

It is not safe to rely upon the spacers and struts of the switch to give proper r.f. connection between the two boards. However it is very convenient for the switch to anchor the two boards mechanically together, and so when the switch has been mounted on PC2 circuit board, the first wafer is separated from the switch body by two $\frac{5}{32}$in spacers and is clamped in position with 6B.A. nuts, with a washer. A single spacer, $\frac{5}{32}$in, is then put on each strut, and then the upper circuit board. Two more $\frac{5}{32}$in spacers are added to each strut, then the upper wafer, and finally this is clamped down with another nut. To enable this to be done conveniently it is suggested that the screw head is first cut off each bolt before assembly. The circuit boards will be separated by $1\frac{1}{4}$in, the wafers by $\frac{3}{4}$in approximately, and the spindle need not be shortened; about $\frac{7}{8}$in of spindle stands proud (see Fig. 10).

Screening

There is a certain amount of "radiation" from the line timebase generator over a wide band of frequencies, and this can be picked up six feet or more from the receiver. The r.f. unit will normally be within this field, and so screening is necessary to avoid the appearance of three faint vertical strips on the picture. Also, the screening prevents hand-capacitance effects and affords a means of mounting. A suitable box consists of three sides of aluminium sheet with three sides of perforated zinc (which is much cheaper and easier to work than aluminium); details for the marking-out of the sheet metal are given in Fig. 11. The tuner is fixed into the "box" by means of threaded rod, which also serves to mount the tuner in the cabinet.

Winding details for all the r.f. tuner coils are given in Table A, page 62, but it should be observed that since this is an "incremental" type of tuner, with limited channel coverage, the constructor may have to do a little adjustment if the channels he requires are not those provided in the prototype. Channels 1, 3 and 5 are specified here. The primary of T1 is thus 9 turns (Channel 5) padded with two auxiliary inductors (L1, L2) on the switch, each of four turns 28 s.w.g. enamelled copper wire. Similarly the oscillator inductor (L9) consists of four turns 18 s.w.g. tinned copper wire, spaced to $\frac{3}{4}$in length, and padded with L10 and L11, which are each of two turns 28 s.w.g. enamelled

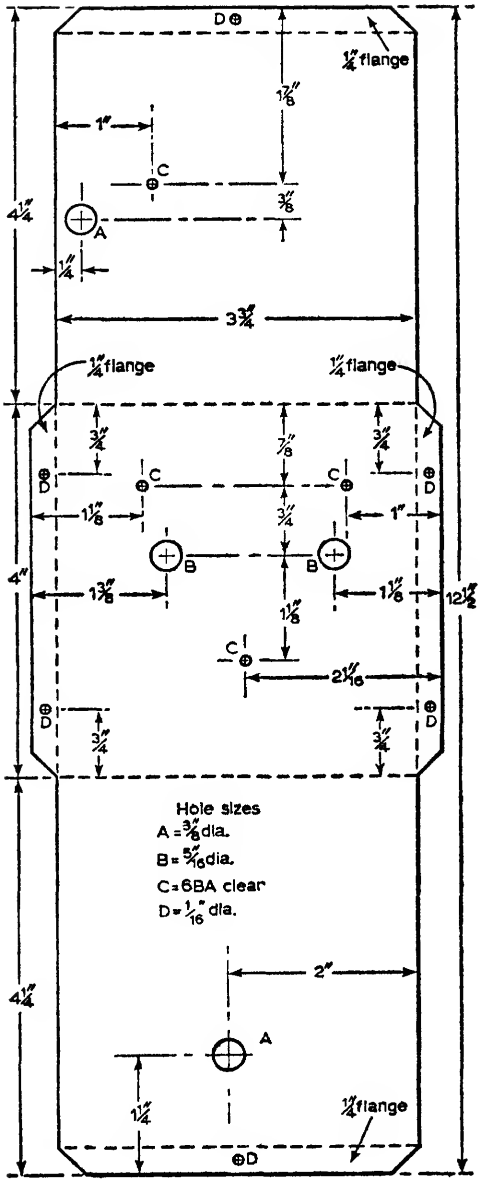


Fig. 11. The screening box for the r.f. tuner unit. A cover (to suit) is required in perforated zinc, and is held in place with self-tapping screws.

copper wire. The fine-tuning capacitor VC1 should cover the channels specified without difficulty, but in case of need the padding inductors may be adjusted by spreading out the turns. To facilitate such an adjustment the padding inductors are first wound on a $\frac{3}{16}$in former and then slipped off. They are strong enough to be self-supporting when soldered to the switch lugs.

The Band III padding inductors consist of $\frac{1}{2}$in of 22 s.w.g. tinned copper wire formed into a hairpin and soldered direct to the centre of the switch lugs. Adjustment is by altering the position of these with respect to the switch by bending them up or down. For the oscillator, smaller hairpins may be needed, and the size may be reduced by squeezing the middle of the hairpin with a pair of pliers until the correct channel coverage is obtained. The prototype was designed to cover Channels 9, 11 and 13.

Setting up the r.f. tuner is hardly practicable until at least the sound i.f. amplifier has been built, and so the "drill" for these adjustments will be detailed later. As a preliminary, however, it is worth while to check that the oscillator functions on all channels—the actual frequency of operation does not matter a great deal at the moment. A very good way of checking the functioning requires negligible apparatus and even less skill. All one needs is an OA70 diode, a multimeter capable of reading 0—50μA or 0—100μA, and a capacitor of approximately 200pF.

The diode is soldered to the two capacitor leads so as to make a loop, and the multimeter leads are connected across the capacitor as close as convenient to the body of the capacitor. This loop is now used to search for r.f. output. It may be looped round a coil, or so placed that a side of the loop runs close to L12, the Band III main inductor. When the tuner is switched on, a d.c. reading should be obtained when the loop is reasonably close to L12.

The same loop can readily be used to detect the presence of unwanted oscillation, and this should be done with the oscillator transistor B− lead unsoldered and the unit switched on, to check that the r.f.-amplifying and frequency-changing circuits are in fact stable. Instability in these is not a problem with this tuner however, and in the prototype these particular circuits could not be induced to oscillate unless something gross were perpetrated, such as touching a collector lead with the finger.

I.F., VIDEO AND AUDIO STAGES

The two i.f. amplifiers required for the 405-line standards are both arranged on a laminate board measuring 12in × 6in. With these amplifiers are associated—on the same board—the sound output and video amplifying stages. Separation of the synchronising signals from the vision signal is accomplished in a separate unit to be described later.

In the prototype, the printed circuit board was finally mounted by the use of metal brackets. These are not shown in the diagrams, but may be made to any reasonable dimensions from 20 s.w.g. aluminium and bolted to the printed circuit in any convenient position.

TABLE A

R.F. Tuner Inductors:

All on formers of Bakelite or polystyrene 0·3in diameter (as obtained with the popular canned assembly)

Band I

Oscillator: L9 (Channel 5) 4 turns 18 s.w.g. bare wire spaced to $\frac{3}{8}$in length. V.H.F. iron dust tuning core (slug)

Padding inductors L10 (Channel 3) L11 (Channel 1). Both, 2 turns 28 s.w.g. enamelled copper wire-wound on $\frac{3}{16}$in dia. drill shank and slipped off. Adjust by spreading the turns.

R.F. Inter-stage: T1 (Channel 5). Primary, 9 turns of 28 s.w.g. enamelled wire close-wound. V.H.F. iron dust tuning core.

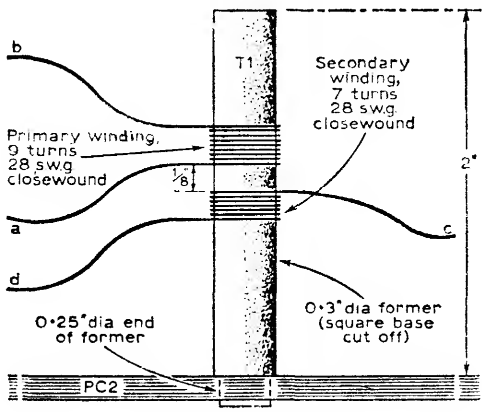
Secondary, 7 turns of 28 s.w.g. enamelled wire close-wound, spacing $\frac{1}{8}$in from primary. V.H.F. iron dust tuning core.

Padding inductors L1 (Channel 3), 5 turns 28 s.w.g. enamelled copper wire-wound on $\frac{3}{16}$in dia. drill shank and slipped off.

L2 (Channel 1), 5 turns wire as above.

L5 (Channel 3), 4 turns. 4 more turns for L6 (Channel 1) as above.

Inductances wired on the switch wafers to be adjusted by spreading the turns. Fixed inductors L9 and T1 (primary and secondary) are tuned by the iron dust cores.



Wiring details of T1

Band III:

Oscillator: L12 (Channel 13) 1$\frac{1}{8}$in length of 22 s.w.g. tinned copper wire, two strands parallel and touching. Adjust by spacing the strands of wire.

Padding inductors L13 (Channel 11) L14 (Channel 9) $\frac{1}{2}$in length of 22 s.w.g. tinned copper wire formed into a "loop" and soldered to centre of switch lugs.

R.F. Inter-stage: T2 (Channel 13). Primary, 3$\frac{3}{4}$ turns 18 s.w.g. tinned copper wire spaced to $\frac{5}{16}$in. Brass tuning core.

Secondary 2$\frac{1}{4}$ turns 18 s.w.g. tinned copper wire spaced to $\frac{3}{16}$in. Brass tuning core.

Note: The space between the primary and secondary windings of T2 must be $\frac{1}{8}$in and must be located so that the primary (connected to S1A) is above the upper circuit board level, and the secondary below.

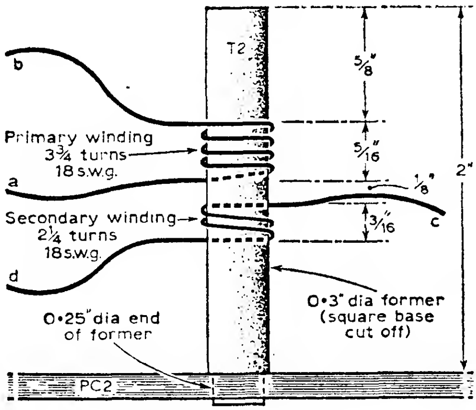
TABLE A (*cont.*)

Band III (*cont.*)

Padding inductors L3, L7 (Channel 11), L4, L8 (Channel 9), $\frac{1}{2}$in length 22 s.w.g. tinned copper wire formed into a "loop" and soldered to the centre of the switch lugs.

(a) oscillator—adjust the positions of these hairpin loops by bending up or down bodily; squeeze centre to shorten wire if necessary.

(b) others—adjust position on switch by bending up or out bodily. In extreme cases adjust length by squeezing centre as above.



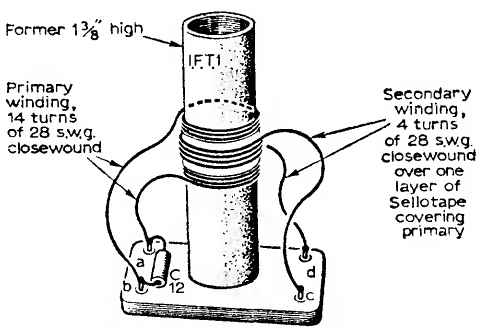
Wiring details of T2

I.F.T.1. Frequency-changer Collector Transformer (Tunes to 27·5Mc/s):

"Short" can assembly ($\frac{3}{4}$in × $\frac{3}{4}$in. × $1\frac{3}{8}$in long).

Primary: 14 turns 28 s.w.g. enamelled copper wire, close-wound.

Secondary: 4 turns 28 s.w.g. enamelled copper wire, close-wound, wound centrally on above, spacing one layer Sellotape. One V.H.F. iron dust tuning core.



Details of I.F.T.1

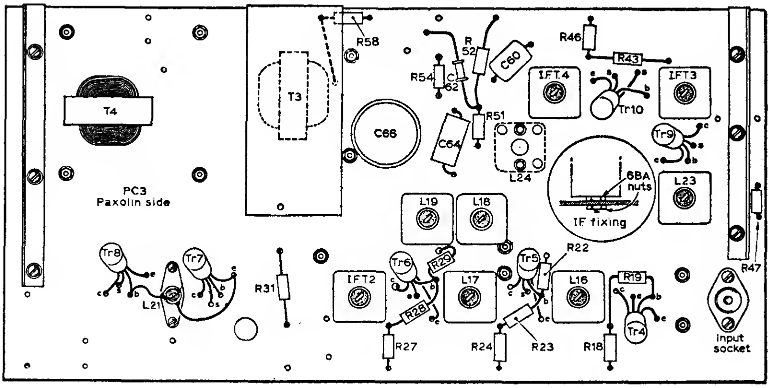


Fig. 12. Layout of the components on the plain side of PC3. T3 is mounted on this side, beneath an aluminium screen. Note the mounting brackets.

In order to obtain good output with thermal stability it is necessary to mount the sound output transistors on heat sinks. These are cut from 20 s.w.g. aluminium sheet 2¼in square and are mounted on the circuit board as shown. They may be earthed if the LFH3 package is used, as the casings of the transistors are not connected to the transistors electrically.

All i.f. screening cans must be properly earthed (to chassis +ve) and if the method for mounting as shown in the inset to Fig. 12 is used, no trouble will be experienced. It will be necessary, however, to use ½in lengths of 6B.A. studding (or 6B.A. grub screws). Bolts can be used, if fitted from the inside of the former.

The Sound I.F. Amplifier

The first sound i.f. amplifier stage is very loosely coupled to the i.f. output from the frequency-changer, and consists of a two-stage configuration of high gain in which five tuned circuits are incorporated. During the design stage it was found not a simple matter to obtain a sufficient degree of coupling in the two double-tuned transformers; if the capacitance used for tuning was such that the correct bandwidth was achieved, the coils themselves became so small physically, that differences in spacing them, hardly detectable to the eye, had marked effects on the degree of coupling achieved. This was, of course, due to the low input resistance of the transistors. Tapping-down was tried but again difficulties of reproducibility appeared. Finally it was realised that by introducing a controlled measure of positive feedback in each stage, the damping could be much decreased. This would enable the tuning capacitance to be decreased, the coil thus physically comprising more turns, and in addition, gain and selectivity would be increased. This was not found to be difficult to arrange over the two stages which were now all that was required.

The prototype receiver uses an OC75 audio pre-amplifier following the detector stage, but gain is so high that the output transistors are well loaded with noise alone, at maximum gain—even though the noise level of this receiver is no more than half that of a valved receiver of equal sensitivity. In fact, the pre-amplifier stage could well be omitted if desired. The facility is left in, however, because even when no picture at all can be resolved sound comes through loud and clear, and this is often all that is necessary for long-distance reception.

Matching of L23 to the input resistance of Tr9 is accomplished by a capacitive tap comprising C47 and C48. Neutralising is by means of a variable air-spaced trimmer, TC2, which is adjusted in the early stages of setting-up. The collector circuit of this transistor is coupled to the base of the next by the double-tuned transformer i.f.t.3, in which coupling is arranged to be about 0·9—a little less than "critical". The secondary is tapped for the base of Tr10 to improve impedance-matching.

A similar transformer couples Tr10 to the detector; the only difference between these transformers is that a tapped secondary winding is unnecessary.

The working-point of the sound i.f. transistors is such that compared with those in the vision i.f. amplifier the input resistance is higher. This is accomplished by a small change in the base bias voltage, caused by using a different value resistance network. Adequate d.c. stability is obtained; the receiver has been tested only to 45°C (113°F) but stability is good. The lowest temperature at which it has been tested is 0°C, and it works quite well at this temperature also.

The detector is of perfectly conventional design, as is the noise limiter. The latter does not appear to be quite so effective with transistors as with valves, but little annoyance is caused by what motor-car ignition interference still "comes through". An i.f. filter is also provided.

The sound amplifying stages are also purely conventional and the output is the usual Class B with driver. About 6dB of negative feedback is applied, with beneficial results, but the driver and output transformers specified will enable some 20dB to be applied without instability if desired. At least 1W output is obtained. The distortion level is some 2–3% with 6dB negative feedback, and would decrease considerably with more. Plenty of gain is available should this be desired. The reader will observe that the writer is not a "hi-fi" enthusiast!

A.F. Transformer

It is a matter of considerable disappointment that Messrs. Gilson have discontinued the supply of the very excellent transformers originally specified for the audio stages. Radiospares transformers have been quoted as alternatives, but the makers do not claim that these can "take" more than about 750mW, and when "pushed" a little can cause poor quality reproduction. However, suitable transformers can be wound by hand.

The driver transformer can still be the Radiospares item, but the output transformer should be of rather more substantial size, of primary to secondary ratio about 3·3+3·3:1 for 3Ω speaker, tightly coupled, and of d.c. secondary resistance below 0·2Ω. Primary inductance should be about 0·5 to 1H, or bass response will suffer and magnetising current will be excessive.

Extra negative feedback can be used with the Radiospares output transformer, however, to improve output and diminish distortion, and this is effected by connecting a 100μF capacitor and 5·6kΩ resistor in series, and connecting these between one "side" of the loudspeaker voice coil and the collector of Tr12 (see R110 and C106 in Fig. 13).

Care should be taken to observe capacitor polarity, and to avoid positive feedback by the correct selection of the loudspeaker lead to which connection is made. This gives relatively heavy feedback, but there is ample gain in the sound receiver to overcome the resulting drop in amplification.

In addition, it may improve the noise level to use a selected OC44 instead of an OC75 as Tr11, reducing collector current to about 0·3mA by adjustment of bias.

The sound section of the receiver is set up in the usual way, using a signal generator tuned to 38·15Mc/s. The neutralising capacitors TC2 and TC3 are both set at minimum to begin with. As the tuned circuits are brought into alignment these should be adjusted upwards in value, until, with the receiver fully tuned it is on the verge of oscillation. Then the neutralising capacitors should be unscrewed a little—about 10% capacitance reduction—and the tuning adjusted again. This should be repeated until reasonable noise output is obtained with the volume control at maximum and no signal input. The noise current in a 3Ω speaker should be about 10mA to 20mA r.m.s. (as measured with a G.E.C. Selectest meter set to the 75mA a.c. range—connected in series with loudspeaker and output transformer).

The Vision I.F. Amplifier

It was in the vision i.f. amplifier that serious problems began to arise in the development of the receiver. The "tailoring" of the response curve itself presented no great difficulty, although at an early stage it was realised that double-tuned transformers would raise problems. There are many means of accomplishing the inter-stage coupling; apart from tapped secondary and separately-damped primary to obtain the proper bandwidth, a combination of tuned transformer with π-coupling is possible. This had already been used in the r.f. tuner, and was not difficult to adjust, but in fact it was yet another adjustment to carry out and to be avoided if possible. Besides, the possibility of conversion for 625-line transmissions had to be kept in mind, and if the same inductors could be made to do duty for the present and the future, so much the better as this would ease the conversion task.

It was calculated that a minimum of five tuned circuits would be necessary to obtain an appropriate shape of response curve over the whole receiver, and that because of the bandwidth required the gain would be such as to

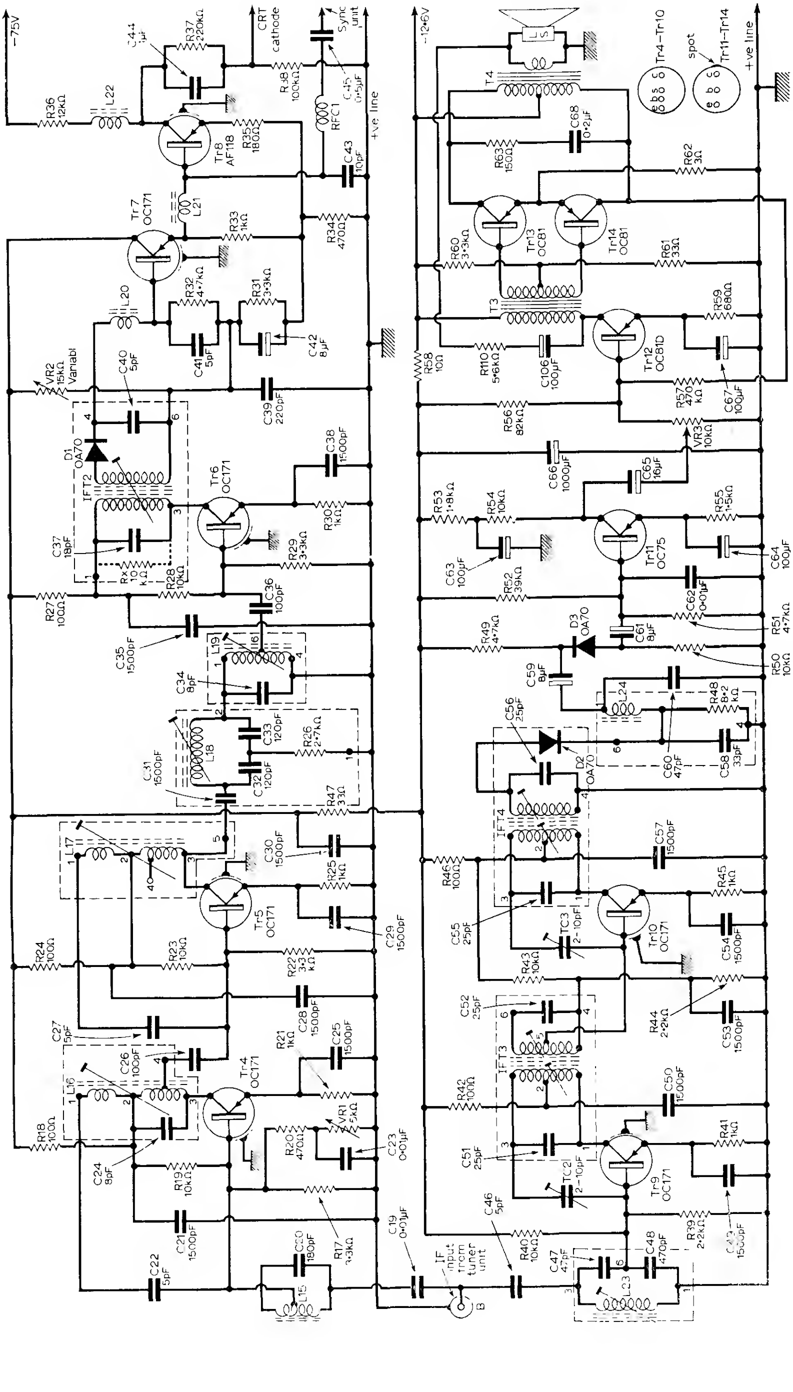


Fig. 13. The circuit diagram of the i.f., video and audio stages. Resistor Rx (contained within the can of i.f.t.2) is a 10kΩ 5% ¼W damping resistor.

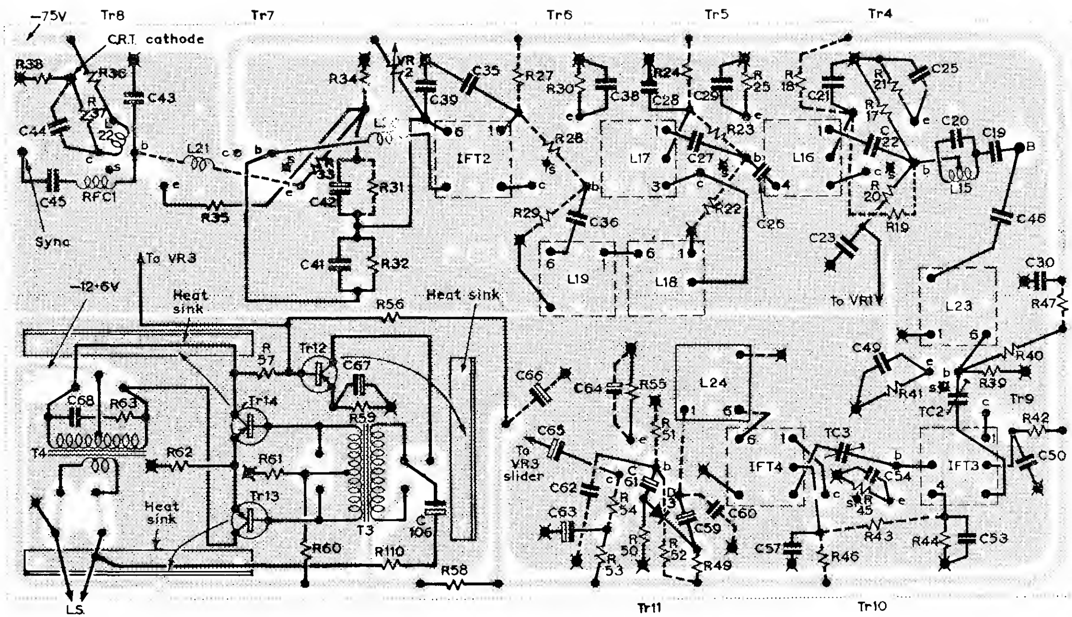


Fig. 14. The printed board PC3 (copper parts are shown shaded). The components are mounted on this *side of the board,* except *those shown dotted* and *T3 and T4 (see Fig. 12). Drilling dimensions of the i.f. transformers should be checked against individual components. The paxolin should be allowed to extend about $\frac{3}{8}$in at either end of the above diagram to allow for mounting-brackets. All points marked "X" are soldered connections to the positive foil. R110 and C106 may be wired as in this diagram, or as in Fig. 13.*

need three i.f. stages. With the frequency-changer collector circuit already decided, this left four more; so that if the experimental and theoretical work had been done properly (as was hoped!) the right response would be achieved if one extra tuned circuit could be fitted in somewhere.

It was when the problem of the sound rejector was being mulled over that one possible solution occurred to the writer. As will be realised, the low impedance at which transistors operate implies that sound traps will have to work at relatively low impedance, and this usually involves tapping well down on a coil, especially if it is in a base circuit. When a coil is already down to four turns or so, an appropriate tap is hard to secure, and in any case coupling is then so low that the resulting rejector has little effect. The way out seemed to be to raise the impedance level at which rejection of the sound i.f. would take place, and this involved fitting the rejector in a collector circuit, where the impedance level is relatively high.

An infinite-rejection type of circuit proved to be very effective, and when used to couple a collector tuned circuit to a base tuned circuit not only resulted in a maximum rejection of some 50dB (using 5% preferred-value capacitors and resistor) but allowed the extra tuned circuit to be fitted in nicely. In fact, to obtain the necessary bandwidth in the rejector, its rejection capability had to be reduced to some 38dB, which is not enough; so a further rejector is used, in the input circuit to the amplifier, of the simple tapped tuned-circuit variety already mentioned. The latter gives about 10–12dB only, but the total rejection in the receiver thus amounts to better than 47dB, which is plenty.

The arrangement finally decided upon is seen in Fig. 14 at L17, L18 and L19 with associated components. The rejector is placed at this point in the receiver, because to put it in the detector stage, where the signal is of greater amplitude, would invite cross-modulation between sound and vision signals (which could never be removed) because of non-linearities in the working characteristics of the transistors. To put it in an earlier stage would have interfered with the layout envisaged for the 625-line conversion, but had cross-modulation become evident in its present position this could not have been avoided. In fact, cross-modulation is negligible with the present circuit arrangement, probably because the input rejector L15 attenuates the sound i.f. signal appreciably and eases matters to some extent. Fixed neutralisation is practicable and convenient in the vision i.f. amplifier, and no adjustment is called for.

Conventional detection is employed, and the diode D1 is followed by a video amplifier in the usual way. In order to obtain effective detection a moderate value of diode load resistor is required, and this means that the input resistance of the amplifying stage should be at least equal to the diode load resistance required. However, it is much less when used in the common emitter configuration, and this is therefore an impracticable arrangement.

Consequently, the detector is followed by Tr7, a transistor in the common collector arrangement, which provides for a much higher input resistance.

Its output resistance is of course very low, but matches well enough into the base of the second video amplifier Tr8 which is arranged to deliver the necessary voltage drive to the cathode-ray tube.

Tr8 is a high-voltage high-frequency transistor. Because of its low collector capacitance, and the low input capacitance of the cathode-ray tube, a relatively high value of collector load resistance may be used provided high-frequency compensation is used. In this circuit, the compensation is wholly in the collector circuit and consists of an inductance L22 in series with the load resistor. This is perhaps not the best system to employ, but it suffices for the 405-line transmissions which use a video signal of 3Mc/s bandwidth.

Emitter equalisation is a possibility, but if the process is carried too far the input resistance of the transistor may become negative, and oscillation will probably result. If desired, a 100pF capacitor across R35 may be tried, but is not really necessary as the 3Mc/s bars on the test card can be resolved with the present arrangement.

The network C44, R37, R38 serves to reduce the d.c. potential of the c.r.t. cathode feed to such a level that proper control of brightness can be obtained. C44 must be arranged so as to have the least possible capacitance to chassis to avoid high-frequency losses.

Gain control of the i.f. amplifier is necessary, to enable the contrast of the picture to be altered. The simple network VR1, C23, R20 in the base circuit of the first vision stage allows the working-point of the transistor to be altered, and sufficient attenuation is obtained in this way to cover normal requirements. In areas of very great signal strength it may be useful to include a precisely similar network in the base circuit of the r.f. amplifier in the tuner unit. The control exercised is "d.c." and therefore any desired length of lead to R20 may be used, and this applies if an r.f. sensitivity control is also employed. Negligible change of bandwidth occurs when either control is used.

The vision i.f. amplifier is readily set up by applying to the input socket the modulated output of a signal generator and tuning each inductor in turn to the frequency given in the table of inductances (Table B, page 72). To check the output, a pair of headphones may be connected (in series with a 0·5μF capacitor) from the base of Tr8 to "chassis". For this alignment, like that of the sound amplifier, a battery supply of 12·6V is the simplest method —either four cycle headlamp cells or a small 12V accumulator. The total current consumption of tuner and both i.f. amplifiers is only about 25mA. Of course the -75V supply to the video output transistor is not yet connected.

The last step in alignment procedure is to re-tune the signal generator to the point where maximum sound output is received on the *sound* receiver, namely 38·15Mc/s. The volume control should then be operated so as to cut off the sound completely, and if this cannot be achieved, the loudspeaker should be disconnected and replaced by a 3Ω resistor. (When listening on the headphones to a very weak signal, extraneous noises can be very troublesome.) Attach the headphones across C43 (still in series with a 0·5μF capa-

citor) and rotate the core of L18 until the sound disappears almost completely; finish off the process by increasing the signal generator output and rotating the core of L15 for minimum sound. The minimum obtained is quite sharp, and great attenuation is achieved.

The whole receiver is now aligned by plugging in the r.f. tuner and connecting power supplies. Set the signal generator to 37·5Mc/s and bring the output lead near the emitter of the frequency-changer transistor Tr2. Tune the i.f. transformer i.f.t.1 for maximum sound as heard in the headphones still connected to the vision amplifier. For the alignment frequencies of the other vision i.f. tuned stages, see Table B.

When this process has been carried out, all the cores so far adjusted should be locked with a suitable locking compound. If none is available Plasticene makes a good substitute and is less messy than grease.

The alignment of the r.f. circuits is inherently a simple process, but may require some care. The first inductor to be trimmed must always be the basic one; the padding inductors are aligned in turn, step by step round the switch, until all are done. It is also important to align the Band I elements first, as because of proximity effects their setting is found to have a small effect on the Band III elements. On each band, work from the highest frequency to the lowest. For each channel, the r.f. setting of the signal generator should be to mid-channel; for example Channel 1 requires a signal generator setting of 43·5Mc/s.

Last of all, connect an aerial to the tuner unit, select the channel required and adjust the volume control. If all has gone well, the sound should be heard very clearly; if not a slight adjustment of the fine-tuning capacitor will bring it in.

Vision I.F. Response

Fig. 15 shows the response curve of the vision i.f. amplifier. Because of the vestigial-sideband nature of the transmissions the 3Mc/s bandwidth extends from 6dB down at the low-frequency band edge to 3dB down at the high-frequency edge. This is seen to be achieved very closely.

It will also be noted that the response at 41Mc/s is better than 25dB down. This ensures that when Channel 1 is being tuned, the i.f. amplifier does not become unstable because of direct response to the signal on the sound or vision frequencies.

The vision signal can be heard on the headphones if the sound is tuned in correctly, but the actual appearance of the signal at C43 will be of interest. This amounts to half a volt or so, and can be applied to the Y-input of an oscilloscope.

It is useful at this stage to use a dry battery supply to the video output transistor, to make the final adjustment to the receiver. The $67\frac{1}{2}$V Ever Ready type B101 or equivalent is convenient, and if its + terminal is connected to the −12·6V rail of the receiver, −80V will be available. This is

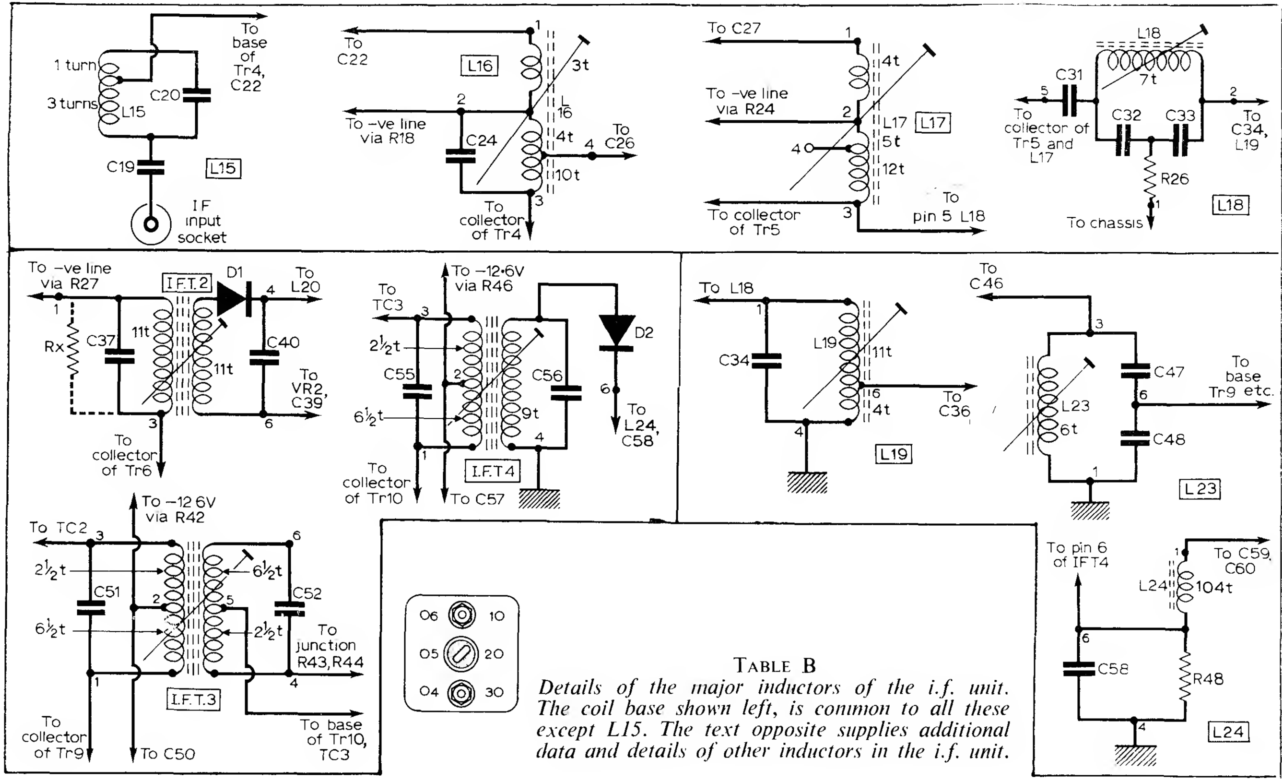


TABLE B

Details of the major inductors of the i.f. unit. The coil base shown left, is common to all these except L15. The text opposite supplies additional data and details of other inductors in the i.f. unit.

TABLE B

I.F. Amplifier Inductors:

All wound on bakelite or polystyrene 0·3in diameter formers, as obtained with the popular canned assembly. Coils wound with the finer s.w.g.s may be fixed in position with polystyrene cement

L15 Sound I.F. Rejector:

Short coil former, no screening can. 4 turns 28 s.w.g. enamelled wire tapped 1 turn from end, turns spaced by wire diameter. Winding shunted by C20. Single v.h.f. (purple) dust core

L16 Vision I.F. Inductor:

"Short" *can* assembly. 17 turns 28 s.w.g. enamelled wire, taps at 10 and 14 turns from "collector" end. Single layer close-wound. C24 included inside can. Single v.h.f. dust core. Tunes to 37·5Mc/s

L17 Vision I.F. Inductor:

As for L16, but taps are at 12 and 17 turns instead (from "collector" end). No internal capacitors fitted. Single v.h.f. dust core. Tunes at 37·5Mc/s. Tap at tag 4 is for future 625-line conversion

L18 Sound I.F. Rejector:

"Short" *can* assembly. 7 turns 24 s.w.g. enamelled wire, single layer, spaced by wire diameter. C31, C32, C33 and R26 fitted inside can. Single v.h.f. dust core. Tunes to 38·15Mc/s

L19 Vision I.F. Inductor:

"Short" *can* assembly. 15 turns 28 s.w.g. enamelled wire, single layer, close-wound. Tapped at 4 turns from "earth" end. C34 fitted inside can. Single v.h.f. dust core. Tunes to 36·5Mc/s

L20, L21 Vision I.F. Rejector:

"Short" coil former, no screening can. 110 turns 40 s.w.g. enamelled wire, close-wound single layer. Standard dust core—set to maximum inductance fully within coil winding—no further adjustment necessary

RFC1:

As L20, L21—no dust core. Should be mounted at right angles to L22

L22 Video Correction Inductor:

"Short" coil former, no screening can. 200 turns 40 s.w.g. enamelled wire, in three layers of 70, 68 and 62 turns respectively. Each layer insulated by 1 turn Sellotape. Standard dust core—set to maximum inductance

L23 Sound I.F. Inductor:

"Short" *can* assembly. 6 turns 28 s.w.g. enamelled wire, spaced by wire diameter. Single v.h.f. dust core. C47, C48 fitted inside can. Tunes to 38·15Mc/s

L24 Sound I.F. Rejector:

"Short" *can* assembly. 104 turns 40 s.w.g. enamelled wire, single layer. Standard dust core, set to maximum inductance. C60, C58 and R48 also fitted inside can

I.F.T.2 Vision Detector Transformer:

"Short" *can* assembly. *Bifilar* wound, primary and secondary interwound. 11 + 11 turns 28 s.w.g. enamelled wire, close-wound, single layer. Rx, a 10kΩ damping resistor. C37, C40 and D1 are fitted inside can. Single v.h.f. dust core. Tunes to 35·25Mc/s

I.F.T.3, I.F.T.4 Sound I.F. Transformer:

"Long" *can* assembly ($\frac{3}{4}$in × $\frac{3}{4}$in × $2\frac{1}{4}$in)

Primary—9 turns 28 s.w.g. enamelled wire, spaced by wire diameter, single layer, tapped at $2\frac{1}{2}$ turns from "TC" end (inner) for negative supply connection. Single v.h.f. dust core

Secondary—as primary, tap at $2\frac{1}{2}$ turns from "R43/R44" end (inner) for Tr10 base. Single v.h.f. dust core (I.F.T.4 has no tap on secondary for base connection)

Primaries and secondaries are wound side by side on the formers but are spaced $\frac{1}{4}$in between windings

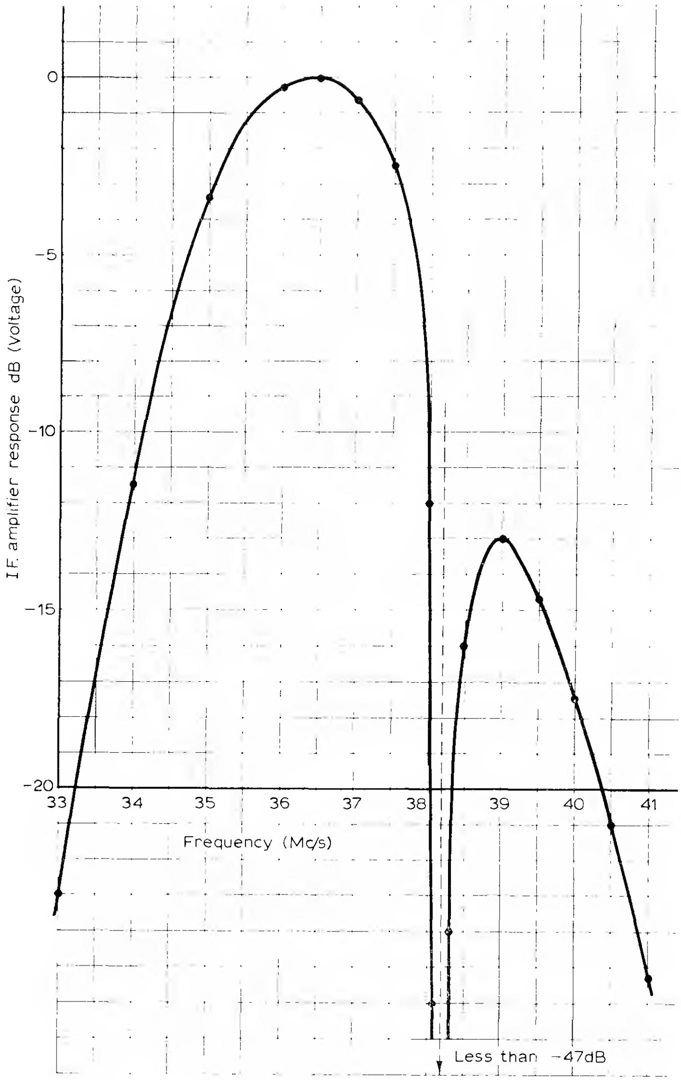


Fig. 15. The response-curve of the vision i.f. amplifier.

quite safe to use. A milliameter should be connected in series, and VR2 adjusted until the transistor draws 4mA (including the bleed current through R37 and R38).

The inductor L22 is best adjusted later on, when the Test Card is being displayed.

THE POWER UNIT

Hitherto, testing on the units constructed has been feasible with a dry battery, but the units next to be described take much heavier currents and unless a small 12V accumulator is available it may be found inconvenient to proceed. Accordingly, a mains power supply is next detailed conforming to the following specification:

Mains input: 50c/s a.c. between 180V and 240V.

Output: 11 to 18V regulated, adjustable.

Hum level: <25mV at 100c/s.

Circuit: Series stabilised.

Current output: Maximum of 1A.

Regulation at 12·5V output: Zero load 12·50V, at 1A load current 12·52V.

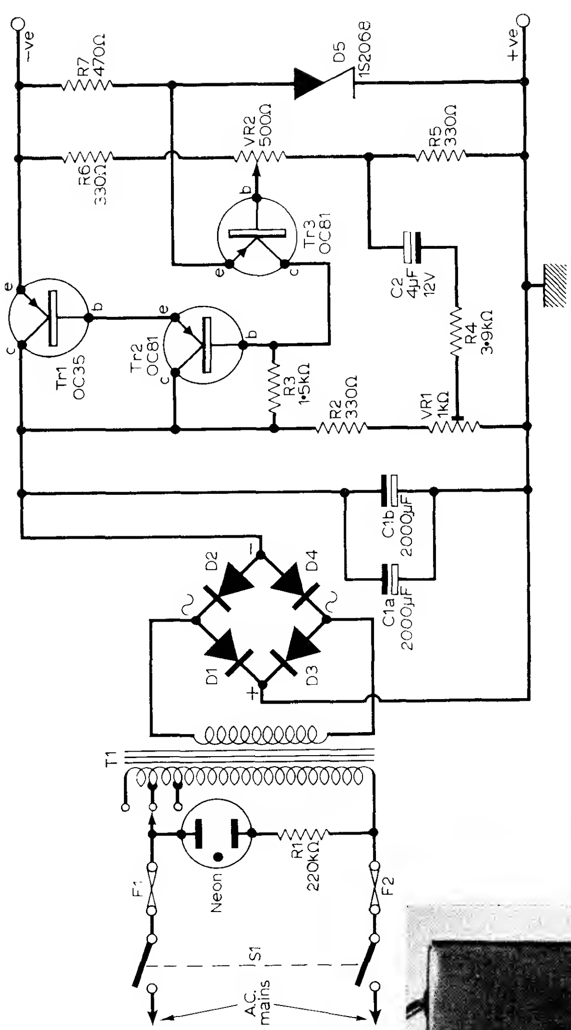
This unit is quite sufficient to supply all the needs of the 405-line receiver, and with a simple modification will cope easily with the extra power required for 625-line working.

Power Supply Circuit

The circuit diagram is shown in Fig. 16, and the simplicity will be evident at first glance. This is made up, in the prototype, on a chassis 6in × 4in, of 18 s.w.g. aluminium. This will support the mains transformer easily as well as the active components. For safety and appearance, it should be totally enclosed, with a fused supply as shown, and provided with a pilot light in the form of a small neon indicator.

The unit consists of a full-wave rectifier assembly comprising four silicon power diodes in bridge formation, fed from a mains transformer. The output is roughly smoothed by at least 4000μF and this forms the input to the regulator circuit. The voltage at the slider of VR2 is compared with the steady voltage across D5, which is a sharp-knee zener diode, and the difference voltage is used to regulate the collector current of Tr3. This is amplified by Tr2, which acts as a current amplifier in the common-collector (or emitter-follower) configuration, and the amplifier current—which forms the base current to the series regulator transistor Tr1—maintains the voltage output at a figure determined by the position of the slider of VR2. The greater the current amplification of the combination Tr2/Tr3, the more perfect the regulation. Since the regulating action also removes hum, it is worthwhile to select Tr2 and Tr3 for maximum gain at 100c/s, although samples taken at random give effective results.

Hum is further balanced out by the network which comprises a 3·9kΩ resistor and 4μF capacitor. With the 405-line receiver the arrangement is correct as it stands in Fig. 16, but where higher currents than 1A have to



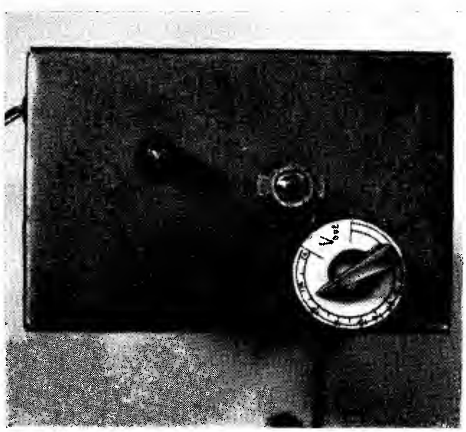


Fig. 16. The circuit diagram of the Olympic II power pack and, left, the appearance of the finished power pack.

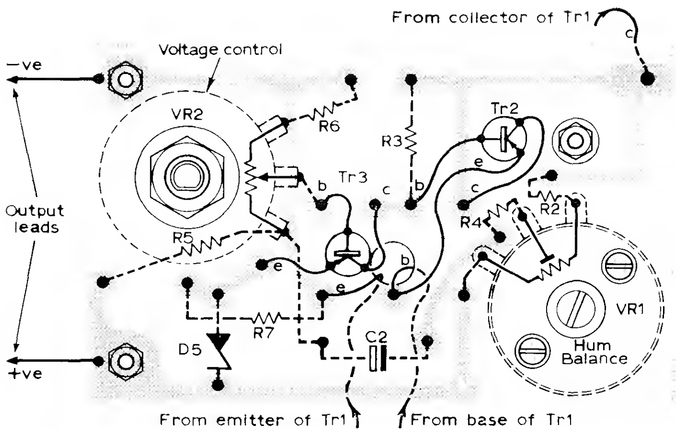


Fig. 17. Details of the printed board of the power pack (copper parts shaded), together with component connections.

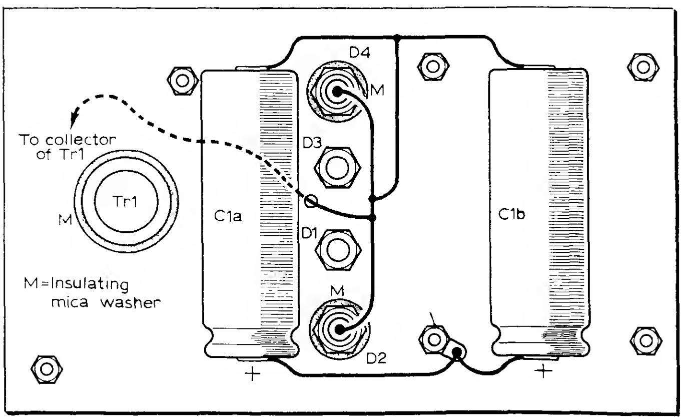


Fig. 18. The components mounted on the underside of the power-pack chassis.

be supplied it may be necessary to change over the positions of VR1 and the 330Ω resistor in series with it. Balance is obtained by putting a 14Ω resistor (made up with resistance wire) across the output terminals, attaching headphones across these terminals also, and rotating the spindle of VR1 for least hum with the output voltage adjusted to 12·4V by means of the slider of VR2.

A Volts Scale

It will be found advantageous to arrange VR2 to be operated by a pointer knob, and to place a volts scale behind it. The scale can be marked in volts, sub-divided into 0·2V units; this enables good setting accuracy.

The unit is immune to mains supply changes of 20%. It is worthwhile however to supply the mains transformer with a tap labelled 210V, when winding, so that if the supply voltage drops below this figure the tap can be changed. This enables the transistors to work under optimum conditions with low supply voltages. However, this refinement is by no means necessary unless the mains supply is normally under 200V. The mains transformer specification is given in Table C, page 80.

It is probable that if the reader possesses a small acid accumulator, he will wish to use the mains unit to charge it from time to time. This can readily be done, but it is highly advisable to put a further rectifier diode, type 1S411(Texas) in series with the supply and the accumulator. If the mains supply is accidentally switched off, there will then be no trouble. Otherwise, if the a.c. supply fails, there will be nothing to prevent the accumulator discharging through the regulator circuit and while the mains series regulator Tr1 may not mind this very much the amplifier transistors will pass relatively heavy current and may be damaged; also the accumulator will discharge rapidly.

Checking

There is no setting-up to be done with this unit; all that is necessary is to check that the open-circuit voltage of the secondary is 16V r.m.s., or a little over, when the primary is connected to a source of 240V. Of the transistors, only the series regulator Tr1 needs a heat sink, and the chassis itself serves admirably in this capacity. It requires an insulating kit, as the collector is electrically attached to the casing. The amplifier transistors need no extra cooling, although the rectifier diodes all need a heat sink also. Again the chassis is suitable for this purpose. It will be seen from the circuit diagram that two of these diodes are connected direct to the chassis, but the remaining two require to be insulated from it with the usual mica washer and bushes for the holes. A light smear of silicone grease is a good thing to add between the mica washers and the surfaces they separate, as it assists in the removal of unwanted heat.

Performance and Temperature

In use, the transformer rises a little in temperature, but not so much as to cause interference with the operation of the unit. The transformer has been designed to cope with the much heavier currents required with 625-line operation. Its temperature will be found to rise much more when delivering the 2A needed in the latter circumstances, and it is therefore well worthwhile

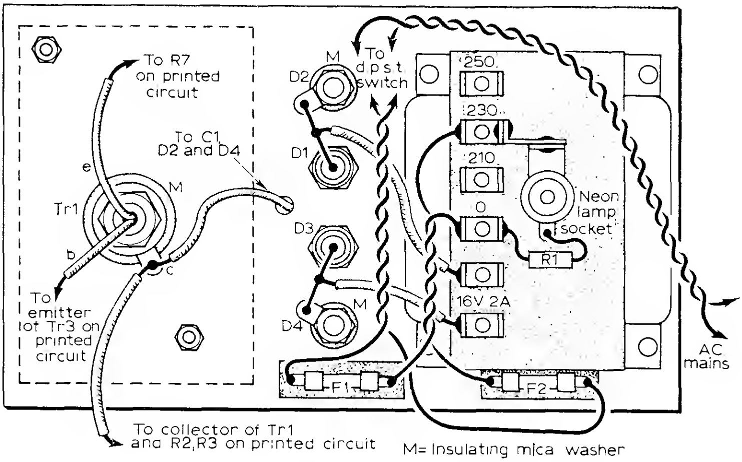


Fig. 19. Layout and wiring details above the power-pack chassis.

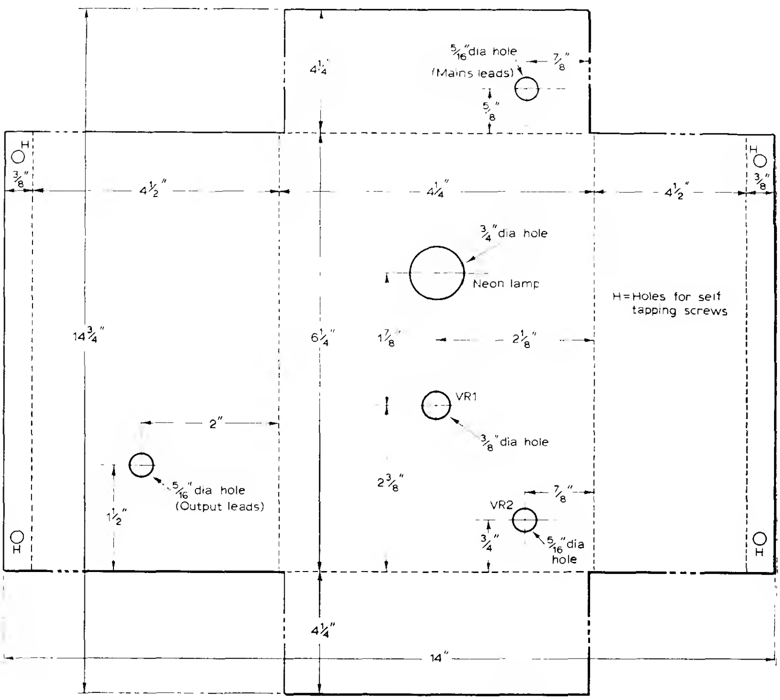


Fig. 20. Dimensions and drilling details of an aluminium case to cover the power pack.

TABLE C

MAINS TRANSFORMER

Core: 1¼in stack of No. 29 M.E.A. laminations Silcor 2, or equivalent No. 29.	*Secondary:* 80 turns 18 s.w.g. enamelled copper wire, in four layers, no interleaving
Primary: 1200 turns 32 s.w.g. enamelled copper wire in eleven layers interleaved 0·001in paper. Tap if desired at 200 turns and use 1000 turns for supply under 200V	*Primary and Secondary:* Separated by two turns empire cloth or insulating tape in equivalent thickness *Outer Covering:* Insulating tape. (The windings are at 5 turns/volt)

to mount the transformer on a heat-insulating platform above the chassis. The chassis and its cover should be well ventilated, and should be matt-painted on the outside to assist in removing heat. The rectifier diodes and Tr1 should be mounted at the opposite end of the chassis from the transformer. The two smoothing capacitors are quite large physically, and if the chassis is made 1¼in deep they can be put underneath it conveniently.

Integral or Separate

This mains unit is intended to be a separate bench-type supply, as it is fairly large. If desired, however, it can be accommodated in the receiver itself.

Line and Field Generators

The design of both line and field scan generators is in practice dependent on what is commercially available by way of the scanning yoke, since this is a highly specialised component whose construction is beyond the scope of the average constructor. Fortunately a suitable yoke is obtainable from Electronic Components Ltd. (Elac), 33 The Hale, Tottenham, N.17. This consists of a ferrite ring wound with 37Ω frame scan coils and nominally 140μH line scan coils. With this can be obtained a suitable e.h.t. transformer and a field output choke. These components will also be required and their use will be described.

Output Stage

The line scan circuit is shown in Fig. 21 and it will be seen that the output stage uses a power transistor arranged as a switching device; in the collector circuit, the line-scan coils are connected, and in parallel with these, the primary winding of the e.h.t. transformer. How current divides between

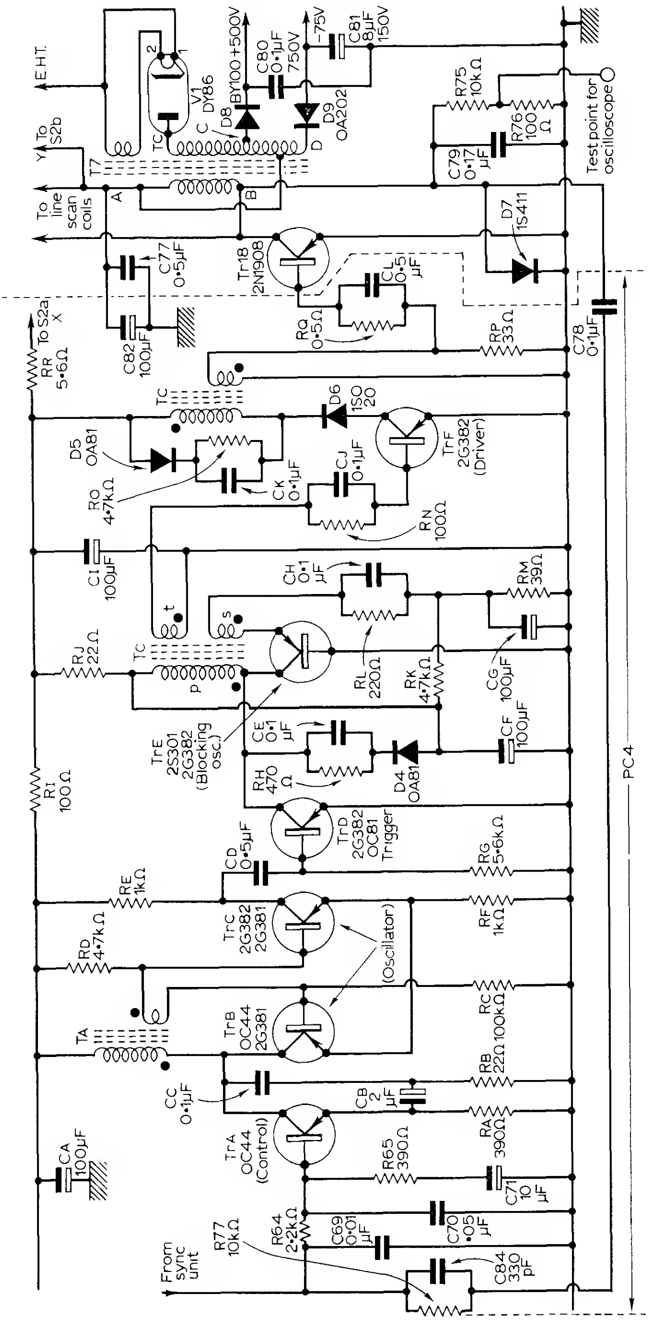


Fig. 21. The line timebase and associated circuitry. The dots on the transformer windings indicated the start of each winding. C79 is a close-tolerance component made up of 0·15µF and 0·02µF capacitors in parallel.

the e.h.t. transformer and the scan coils depends partly on their relative resistances but much more on their relative inductances. The e.h.t. transformer inductance is nominally 1·5mH and thus the parallel total inductance is 125μH approximately. The scan coils thus get very nearly 90% of the collector current, while the transformer receives 10%—quite sufficient for the needs of the cathode ray tube and the video amplifying stage when transformed to the appropriate voltages. The d.c. resistances of both yoke and transformer are very small and can be neglected in calculating feed currents and voltages, although the resistance of the yoke—and the output transistor—introduces a small non-linearity into the scan which has to be corrected.

The output transistor Tr18 is a Texas Instruments *pnp* power device capable of switching up to 20A collector current. Its collector-to-base-breakdown voltage rating is −130V and its frequency-handling capacity extends to several Mc/s. The type is 2N1908 and may be had on order with a delivery delay of a week or two. As might be expected, this very high quality transistor is somewhat expensive, but this need occasion no alarm, since under normal working it would outlast scores, if not hundreds, of line output vacuum tubes. However, it is most important that this transistor should be worked within—if close to—its limits, as all transistors are readily destroyed by improper use. The makers do not specify an absolute maximum voltage rating because exceeding breakdown voltages does not permanently damage transistors as long as other maximum ratings are not exceeded. In particular, the instantaneous current rating is important because if this is exceeded and the voltage is high the power dissipation may be very great locally within the transistor. Suppose that while the collector-to-emitter voltage were momentarily at 100, the emitter-collector current were 20A, this would mean that 2,000W would be produced, which is well above the transistor rating of 150W maximum. Even though this lasted for only 1% of the "duty cycle"—giving an average dissipation of 20W—the heavy current produced instantaneously would cause a large local magnetic field within the transistor; and this, in a semiconductor material, would be likely to result in a "pinch" effect, constraining the current to flow in a very narrow channel. The result might well be to increase the *local* dissipation within the germanium slice to destruction point.

It may be remarked that no experiments have been carried out on these lines and care has been taken that ratings are not exceeded. If this design is followed with care no harm is likely to come to the transistor, but the constructor must always be aware of the physical nature of the situation and will do well to remind himself of the mathematics, too, if possible. Several pounds per microsecond is above most people's income!

The design problem included the assessment not only of the scan coil current required (which is just over 7A) but also of the flyback characteristics of the assembly on which depends the power dissipated both instantaneously and by way of steady battery supply as well as the peak voltage developed across the transistor. If a battery of fixed voltage is connected in series with

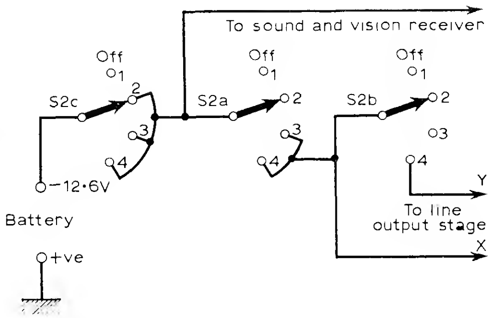


Fig. 22. The function-switch S2. This is shown now only to help illustrate the operation of the line timebase. Full details of S2 appear later.

the scan coil inductance to the collector circuit nothing happens until the transistor is switched on by a negative supply of current to the transistor base. If this current is large enough, the forward resistance from collector to emitter becomes very low—a fraction of an ohm—and current begins to flow in the collector circuit. Because of the inductance of the scan coils the current rises linearly with time until either the transistor is switched off—by means of a heavy positive current in the base circuit—or the transistor is destroyed. The trick is to switch off in good time! At this point, the capacitance across the scan coils, together with their inductance, form a tuned circuit and this begins to oscillate. At the end of the scan, when switch-off occurs, there is a very considerable amount of energy stored in the magnetic field of the yoke, and as this collapses a reverse current is driven into the capacitor, charging it up. The transistor collector now becomes the emitter, since the polarity of the applied voltage is reversed, and is not very efficient at this. Consequently a power diode is connected in parallel with the transistor to pass the reverse current. The end point is reached when the flyback finishes with the collector positive and the parallel capacitor fully charged. Then current again begins to flow into the scan coils as before.

If nothing more were done, several cycles of oscillation would virtually put an end to scanning because of losses. However, if the transistor is switched on by a further negative current in the base circuit the losses are automatically made up by the battery and if this is done regularly each cycle (by a driving waveform) the process repeats indefinitely. Clearly the transistor only needs switching on in each cycle when the *reverse* collector current reaches zero (from a positive value) and begins to go negative again. The first portion of the scan is then due to current passing through the diode—usually about 45% of scan—and the transistor is switched on to provide the remaining 55% of the scan.

Since the capacitor in parallel with the transistor and diode forms part of the oscillatory circuit, the flyback period is just half a cycle of the natural frequency of this circuit. The flyback peak voltage produced depends directly on the speed of flyback and therefore the value of the capacitor is critical. Too low a value causes a higher peak voltage to be developed, while too high

a value lengthens the flyback period and some of the picture is lost. In order to keep the peak voltage to a safe value it is necessary to sacrifice part of the picture, since the line blanking period transmitted is too short. The lost portion is equal to the width of the black-and-white border on the test card and may be neglected. Visible fold-over does not occur because the flyback pulse is used to blank the cathode ray tube and all that is noticed is that a half-inch of picture disappears. Full scan—even a little overscan—is, however, available.

Study of Fig. 23 will reveal that the collector dissipation is limited to a very short period of time, while collector current is falling to zero and the flyback voltage is high. The time the collector current takes to fall to zero (as distinct from the time taken for flyback current to flow into the parallel capacitor) depends on the rate at which the base drive can attract "holes" from the transistor base. To minimise this time, a high-frequency transistor is essential. This is the reason for choosing the 2N1908 device; there are transistors whose voltage rating is higher—much higher—but these are usually relatively low-frequency transistors with a cut-off frequency which is measured only in kc/s. It will be seen also that the "economy diode" also has high dissipation because of the flyback voltage and the current flowing at the same time. For this reason the diode must also have a high voltage rating and be capable of passing a heavy pulse current. The collector-base diode of another 2N1908 would, of course, be ideal but the cost would be prohibitive. Instead a cheap but efficient silicon rectifier diode is needed: the Texas Instruments type 1S411, which has a voltage rating of 200V and a pulse current-carrying capacity of 15A.

The base-drive switching waveform does not need to be the precise rectangular wave shown in Fig. 23. As long as at any moment it is capable of supplying enough current to the base to sustain the collector current required at that same moment, the stage will work satisfactorily. With an output transistor of minimum current gain (15) the peak current of, say, 4A can be ensured with a base current of 4/15 or, say, 0·3A. This is the minimum base drive needed at the end of scan. To switch off the transistor, however, all the stored base charge has to be extracted in about 4μs and this may need as much as 1A reverse drive at the beginning of flyback. Also, since the stored base charge acts as a kind of tank, base drive pulses of less than 4μs duration are unlikely to cause the transistor either to switch on or off (depending on diode polarity). Thus it may be seen that a quite irregular base switching waveform may do the right job provided the irregularities do not exceed certain limits.

The switching waveform can be obtained in a number of ways, among them multivibrators and blocking oscillators. The choice of a blocking oscillator for this application was governed by practical considerations.

Since the original publication of the constructional data of the *Olympic II*, work continued on improvements to what experience had shown to be a reliable receiver. The original circuit for the line-scan unit proved to have a relatively minor defect in that until the blocking oscillator transistor had

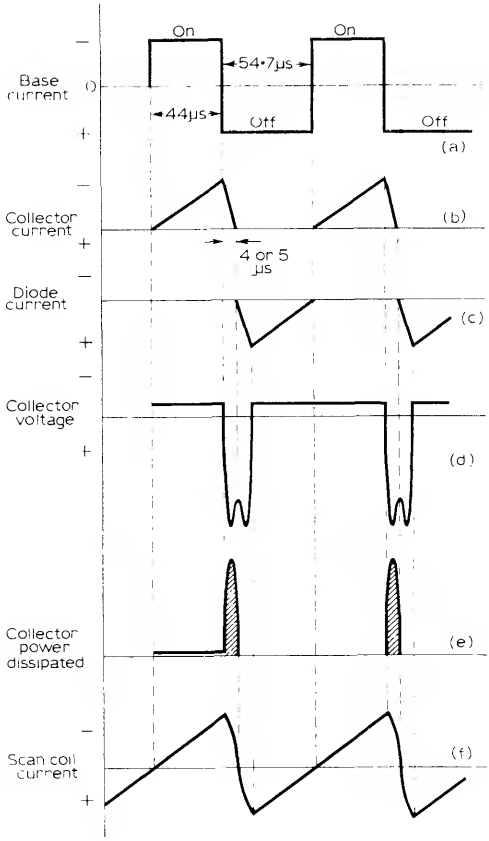


Fig. 23. Waveforms in the line output stage.

reached a stable working temperature there was a slight drift of operation which necessitated the readjustment of the line hold control. After about fifteen minutes operational stability was reached, but this characteristic was considered undesirable and steps were taken to correct the fault.

Readers will be interested to note that this drift should have been correctible perhaps by increasing the gain in the feedback loop. The ability of negative feedback to correct for amplitude distortion is well known, but it may be less widely recognised that if frequency can be related directly to a voltage or current, negative feedback can correct for changes of frequency also. This is what is done in effect when an error signal (an "off-frequency" voltage) is used to afford automatic frequency control in v.h.f./f.m. tuners. However, it proved rather difficult to obtain enough gain in the feedback loop without running into problems of drift in the d.c. amplifier itself. Here, not only was transistor junction temperature one cause, but also variations in the ambient

temperature. Low-leakage silicon transistors might have proved to be the answer to this problem, but these are expensive if of the required quality.

In electronic engineering there is usually more than one way to produce a desired effect, and in fact the straightforward way sometimes turns out to be hopelessly clumsy—especially if it results in complication or the use of highly specialised components. In tackling this problem it proved to be more elegant to return to first principles. Here was an oscillator which tended to drift in frequency; could one be found which was inherently stable? One of the reasons for rejecting the multivibrator as line oscillator had been its voltage and temperature dependence. While the blocking oscillator was better in these respects, and had also the advantage of rather easier 625-line conversion, it was still not good enough. The question seemed to be whether a tuned oscillator—using the highly stable L-C circuit—could fill the bill. It certainly seemed logical to approach the task from this angle.

Since the transistor is essentially a current-operated device, there is no real possibility of being able to couple it to a tuned circuit as lightly as a vacuum tube can be coupled. Unless a transistor circuit of very high input and output impedances could be devised, the transistor would load the tuned circuit heavily, so reducing its Q and so, the inherent frequency-stability. An input emitter-follower might precede the amplifier, and an emitter-follower after the amplifier would complete the circuit. This would have three transistors, and still be inferior to a single pentode.

However, a transistor need not be used as a linear amplifier to maintain oscillation; it can be used as a switch to feed extra energy into a tuned circuit to make good the losses. Using the transistor in this way goes as it were to the other extreme and couples it to the circuit very tightly. However, ideally the transistor is cut off completely or else is hard on in this mode of operation, and so ideally takes the place of a switch either open or closed. It does not (ideally!) load the tuned circuit at all, since it is in series with the tuned circuit, not in parallel. It need hardly be said that this is the ideal case. In practice, the tuned circuit has to supply some power to the transistor base to make it switch. If, however, the transistor circuit is highly regenerative of itself, the tuned circuit need only supply a trigger pulse at the moment of switching; the "duty cycle" will be small and the effect on frequency negligible.

The circuit used is shown in simplified form in Fig. 24. It consists of two transistors arranged in the "long-tailed pair" configuration, familiar to experimenters in the form of the cathode-coupled multivibrator using valves. Instead of a resistor in one collector circuit, there is the tuned circuit, and instead of cross-coupling by means of a capacitor, this is effected by a small winding coupled magnetically to the tuned circuit. In operation, current is switched alternately to one or the other transistor, and hence a rectangular current wave is supplied to the tuned circuit—just as if it were alternately switched mechanically to a source of power.

Readers will note that in Fig. 24 no d.c. bias supplies are provided, but if this be done (by a resistive network attached to one base) and a simple working circuit hooked up for experiment, the circuit is easy to get going,

and very reliable in operation. It is instructive to connect an oscilloscope across the tuned circuit L1-C1 and to observe the waveform resulting. It is a very pure sine wave, if the Q of the tuned circuit is reasonably high; just the kind of waveform desired by tape-recordists for bias or erase.

For the present purpose, however, the important feature is that in the calculation of frequency, transistor parameters enter into the result to only a very small extent, showing that the frequency is dependent almost entirely on the tuned circuit. Thus, there is here a very stable "clock" mechanism which should be highly suitable for a line oscillator.

If this tuned-circuit waveform were used to drive the scanning stages direct, however, quite a heavy load (transistor-dependent) would be imposed on it, and much would be lost. As it happens, one does not need to do this. If a small resistance is connected in the collector lead of Tr2 the rectangular current pulse through Tr2 will develop a rectangular voltage wave-form across it. This may also be observed readily on an oscilloscope. This pulse could be amplified and used direct to drive the scanning stages. It does suffer however from having a relatively slow rise-time (although probably fast enough) and is of equal "mark" and "space" periods. This would probably do quite well, but the present design is arranged to make matters technically correct. Consequently this pulse is used to trigger a one-shot blocking oscillator using a ferrite component as the feedback transformer, whose pulse duration and rise-time are amenable to separate function design. This in turn drives a driver stage, which operates the output device.

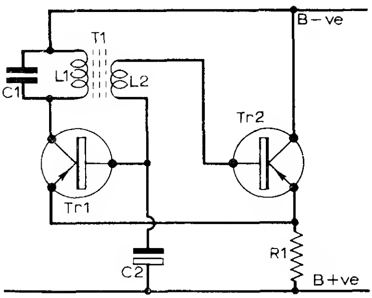
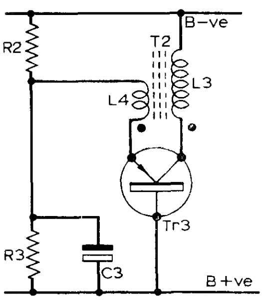
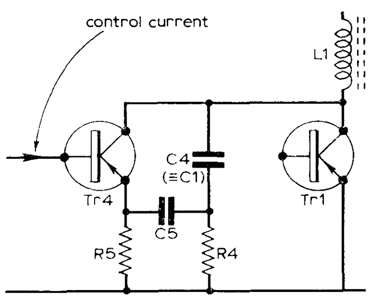


Fig. 24 (left). *Simplified oscillator circuit.*

Fig. 25 (lower left). *Simplified blocking-oscillator circuit.*

Fig. 26 (lower right). *Circuit of the reactance transistor.*





The Blocking Oscillator

The blocking oscillator is shown in simplified form in Fig. 25 and will be seen to be an emitter-coupled blocking oscillator. It is arranged not to be free-running; since the base of Tr3 is held positive to the emitter, the transistor is held in the cut-off state. When a trigger pulse is applied to the emitter (a positive pulse is required) or to the collector (a negative pulse), the base momentarily becomes negative with respect to the emitter and the transistor enters the "active" region of operation. Provided the pulse is of sufficient amplitude to do this, and is of long enough duration, regeneration takes place and the oscillator "fires", producing a pulse whose characteristics depend on the core material of T2 and the nature and disposition of the windings thereon. In the circuit actually used, the rise-time of the pulse is of the order of 0·75μs, and though this performance is degraded by the subsequent driver stage, the pulse edges are very sharp and of very good amplitude—about 1μs rise and fall time. The output transistor is thus switched very rapidly and spends little time in the "active" switching region; thus it dissipates little and safety is assured.

The flywheel sync will be described next. The essential circuit for the sync control is that of the reactance transistor, and is shown in Fig. 26. Considering the tuned circuit of Fig. 24, the capacitor C1 is replaced by C4 and R4 in series. R4 is very small compared with the reactance of C4, and thus a voltage is developed across it which is essentially in quadrature with that across the tuned circuit. This voltage is applied to the emitter of Tr4, via the capacitor C5, and is amplified by Tr4 acting as a grounded-base amplifier. Thus the transistor Tr4 acts as a reactance, and placed across the tuned circuit, can alter the frequency of operation. The amplification of Tr4 is dependent on the working-point of the transistor, and thus if the control current into the base of the transistor changes, so does the frequency of oscillation. The circuit is therefore arranged to utilise the control current from the sync unit to be described later.

The description above should give a reasonably clear idea of the principles used in the line-timebase circuit. See Table D for details of transformers used.

It will be noted that in Fig. 21 the long-tailed pair oscillator does not have an electrolytic capacitor between one base and "chassis"! The use of this would entail the working voltage not being reached for a second or two. Entirely satisfactory results arise from relying only on the intrinsic base-to-emitter capacitance of the transistors, no physical components being required.

Transformer TA

This consists of a Denco unit (in screen) wound with 350 turns of 42 s.w.g. enamelled copper wire (collector), one layer of 0·001in paper, then 75 turns of the same wire for the base connections of Tr2 and Tr3. It may be found difficult to get on all this wire, unless wound very carefully, as the

TABLE D

WINDING DETAILS OF TRANSFORMERS IN THE LINE TIMEBASE

TA

Denco type 9A (in screen) wound with 350 turns of 42 s.w.g. enamelled copper wire (collector winding). Then, one layer of 0·001in paper and 75 turns of 42 s.w.g. enamelled copper wire (for base connections of TrB and TrC). See text for details of alternative numbers of turns and wire-gauge

TB

Core—pair of ferrite cores type FX1238 (Mullard), no gap

Primary: 126 turns 32 s.w.g. enamelled copper wire

Secondary: 63 turns 32 s.w.g. enamelled copper wire, no interleaving insulation

Method—three strands wound together ("trifilar") ends 1, 2, 3; 1′, 2′, 3′

Connect 2′ to 3. Primary is then 2 to 3′, secondary 1 to 1′

Tertiary winding: 60 turns of 32 s.w.g. enamelled copper wire close-wound on top of the primary and secondary windings

Inductances, measured under small-signal conditions at 1kc/s, no d.c. component

Lp = 9·6mH, Ls = 2·5mH

L leakage = 38μH

TC

Core—pair of ferrite cores type FX1239 (Mullard), no gap

Primary: 4 layers, in all 122 turns 26 s.w.g. enamelled copper wire

Secondary: 1 layer, 16 turns 24 s.w.g. enamelled copper wire

Method—2 layers primary, 32 ; 31 turns. 1 layer paper 0·001in. 1 layer secondary 16 turns. 1 layer paper 0·001in. 2 layers primary 30 and 29 turns

Inductances: Lp 4·7mH

Ls 85mH

L leakage 40μH

winding can only be 5/32in in length and the ferrite cup must not chafe it when pushed into position. If this is found hard to comply with, fewer turns may be put on (proportionately in base and collector windings) and the tuning capacitance increased from 0·1μF to perhaps 0·15μF. Alternatively, 44 s.w.g. wire will easily fit the space available. The two "anchors" shown on the printed circuit are for soldering the securing pins of the screening can.

Part of the line-sync circuit is included on the etched circuit board for the line scan generator, since the flyback pulse is used as a sync gate. Also the timing LC circuit radiates and so screening is provided which also serves to screen the main portion of the sync-separation circuit from unwanted pulses. The extra components are shown in the diagram but properly belong to the sync circuit. The need for effective screening also dictates the use, as a panel control, of the d.c. control of line speed available in the sync circuit.

Constructional Notes

Details are shown in Figs. 27–31. For mechanical stability it is probable that it will be desired to mount all the units on a metal chassis. If this is done the best heat-sink for the line output transistor is the chassis itself, on which it may be mounted, using the insulating kit to be ordered with the

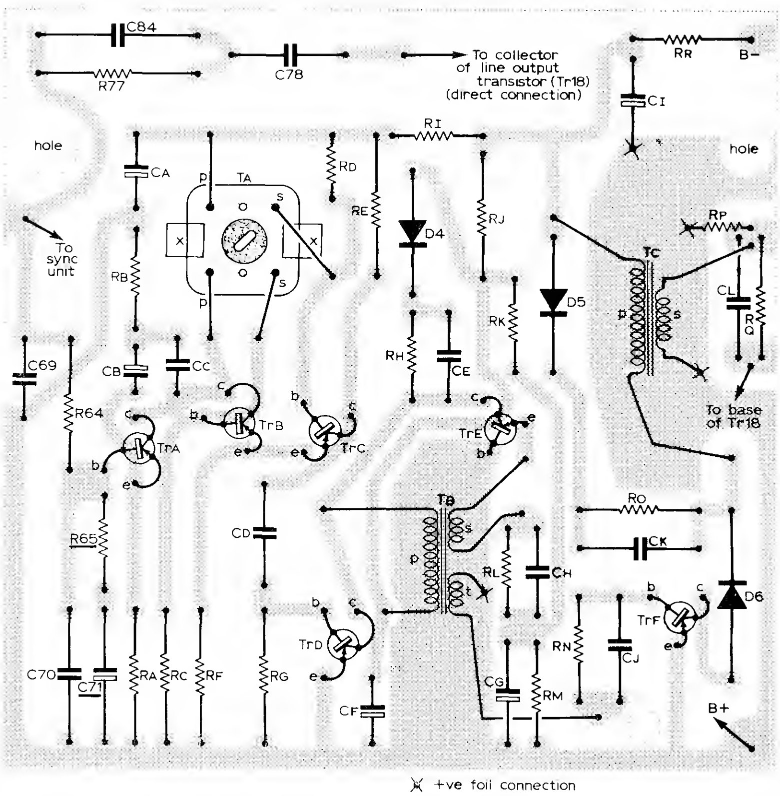


Fig. 27. The printed board PC4 for the line timebase (the copper parts are shown shaded).

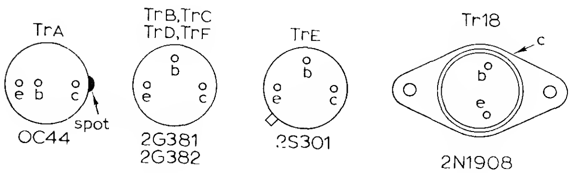


Fig. 28. The base connections of the transistors used in the line timebase.

transistor. The etched circuit board (PC4) for the associated components would then be mounted at a convenient distance above the chassis, using screwed rod, nuts and washers for the purpose, while the e.h.t. transformer T7 would be mounted on the chassis close by, not too far for the e.h.t. lead to reach the c.r.t. in its desired position. The whole assembly would then be covered with a perforated zinc screen secured to the chassis with nuts and bolts.

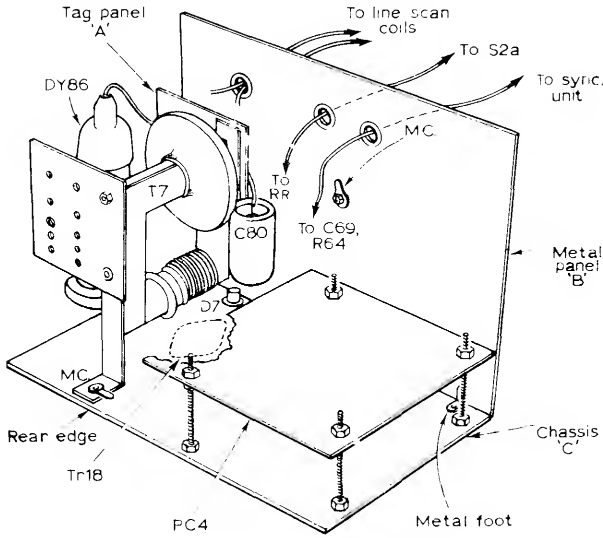


Fig. 29. The complete line timebase unit. This shows how the component parts—the line output transformer, board PC4, chassis "C", etc.—are related when the unit is made up. Note the mounting arrangement of Tr18 (see also Fig. 31) beneath chassis "C" which acts as its heat-sink.

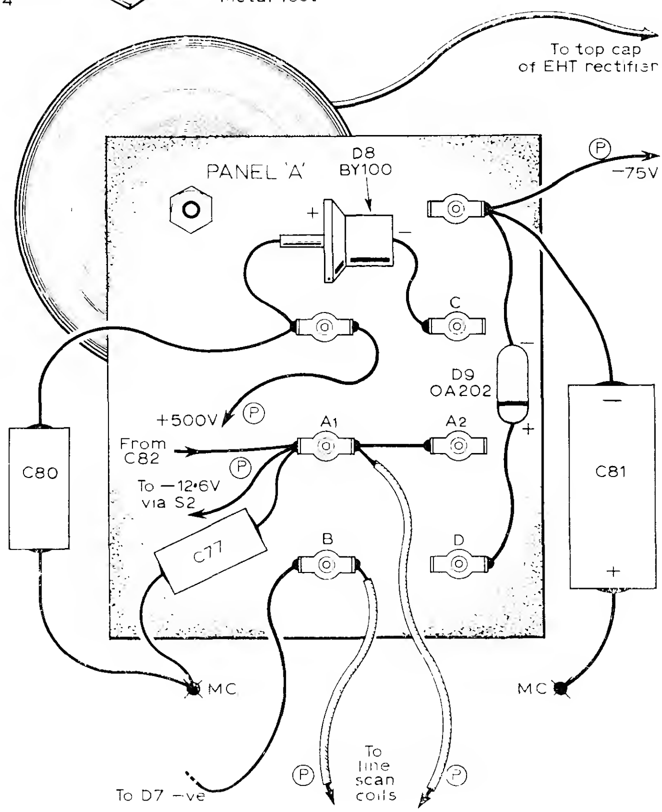


Fig. 30. Component wiring and connections on panel "A" of the transformer T7.

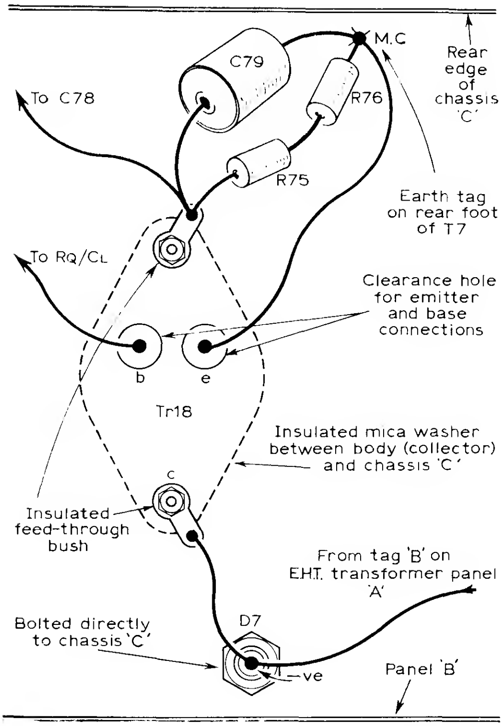


Fig. 31. Details of the wiring of Tr18 and D7.

Layout is entirely non-critical except that the connection from Tr18 to D7 (economy diode) and e.h.t. transformer must be kept as short as possible (Fig. 31). The leads to the scan coils must be of copper braid or 10A flex as they carry heavy currents and negligible resistance must be introduced into the circuit.

Fig. 27 shows a suggested etched circuit board and Fig. 30 the connections to the e.h.t. transformer.

The two tags A1/A2 are connected by a piece of 18 s.w.g. wire, well soldered, and this is the −12·6V connection also. No component should be allowed to come close to the e.h.t. winding or corona discharge is likely; the same applies to the screening box. When the box is fitted, a wise precaution is to fold in some sheet polythene, not too tightly, to ensure that corona discharge is avoided.

Where the negative lead enters the screening box, a 0·5μF capacitor (C77) should be connected by the shortest path to chassis to assist in minimising radiation from the line oscillator.

Heat Sinks

The only transistor requiring a heat sink is the driver TrF, and even this is hardly necessary. About two or three square inches of aluminium sheet will

do quite well. In the prototype such a small sheet was fastened to the clamp of the driver transformer Tc by means of "Araldite" resin. Alternatively, a small finned cooling clip would serve quite well and would need no support.

Setting-up

When completed, setting-up is a simple matter. In preliminary tests, before wiring into the receiver, the frequency should be set up (preferably by using an oscilloscope display, but by ear if no oscilloscope is available), with a 680kΩ resistor wired between the base of TrA and the −12·6V rail. This simulates the actual control current supplied by the sync unit. The frequency of oscillation is adjusted by rotating the core of the transformer TA.

Check with the Thermoammeter

When the unit is installed, make sure that the e.h.t. lead is clear of any objects, insert the DY86 rectifier, and connect the scanning coils. Line and field scan tests and the field scan circuit are shown in Figs. 32–38.

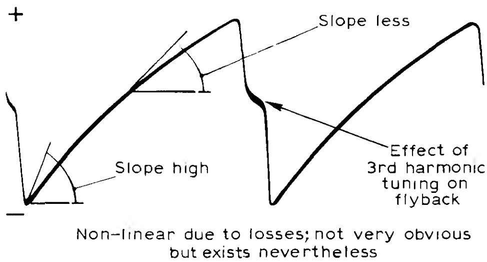


Fig. 32. Waveform of current in the line scan coils.

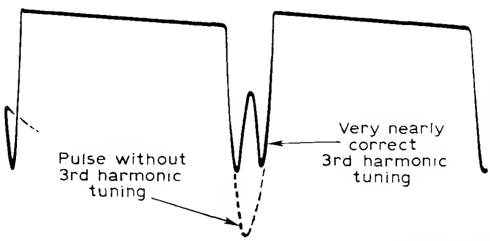


Fig. 33. Line output stage-collector flyback-pulse, showing third-harmonic tuning.

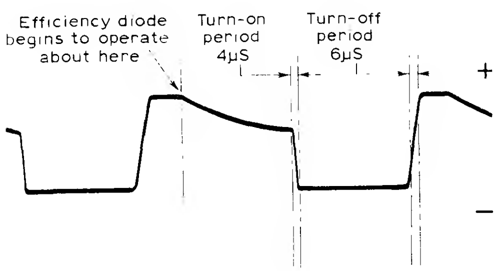


Fig. 34. Waveform at the base of Tr18.

Before bringing the output transistor into circuit the frequency of the oscillator should be adjusted to be nearly correct. The line hold control VR11 (Fig. 39) should be set to the centre of its travel, and with the core of T1 screwed about two-thirds of the way home the potentiometer VR12 (Fig. 39) set for the correct frequency as nearly as possible.

A check should be made with a thermoammeter in series with the base connection of the line output transistor. Connect a 0–3A thermocouple ammeter in series with the lead from the junction of RQ and CL to the base of Tr18. Switch on the negative supply, and if there is a small reading only on the meter, reverse the connections of the secondary winding of T6. The reading should be about 0·5A, perhaps 0·6A. If it less than 0·45A, reduce the value of the base series resistor a little; if more than 0·6A the value of the resistor can be increased a little although this is hardly necessary and a "high" reading is to be preferred.

If an oscilloscope is available a good rectangular wave should be obtained between the base and "chassis" with fast rise time. If the rise time exceeds 2·5μs, something is wrong and the output transistor should not be switched on until the trouble has been found. If no oscilloscope is to hand, put a 2Ω variable resistor in series with the battery lead to the scan coils, and an ammeter in series with the battery supply to the stage. On switching on, the

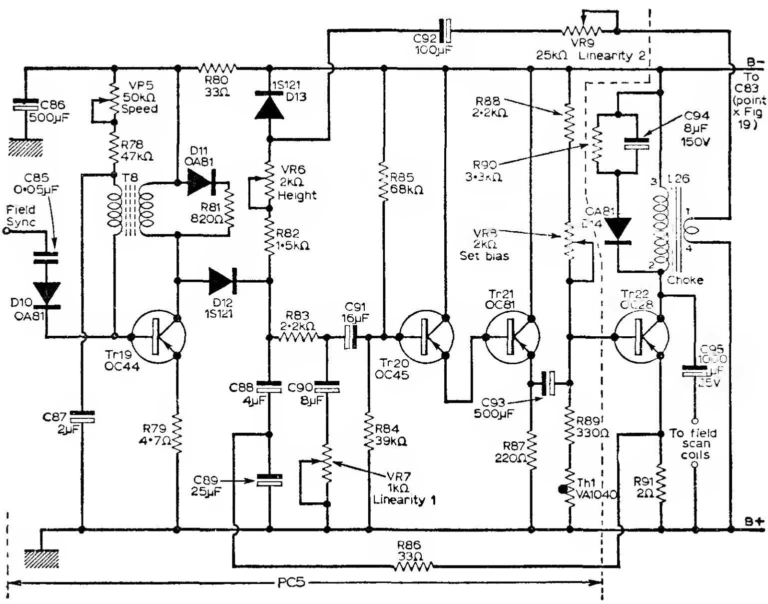


Fig. 35. The circuit of the field-scan generator.

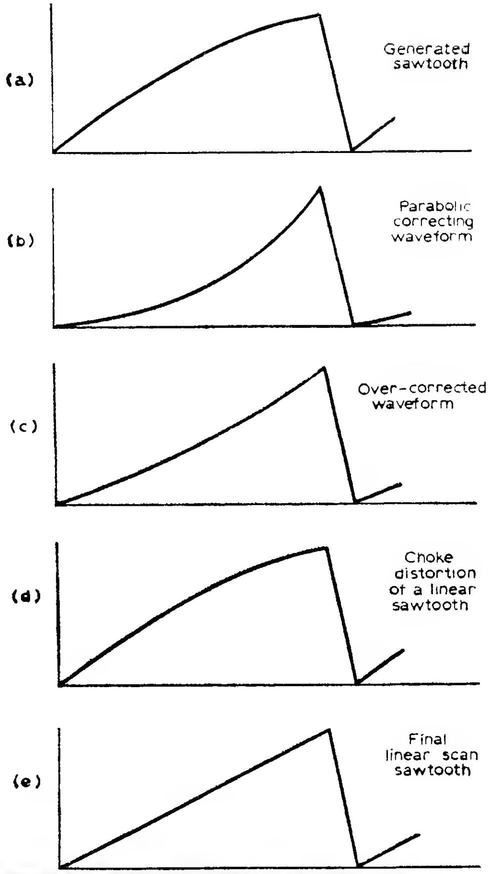
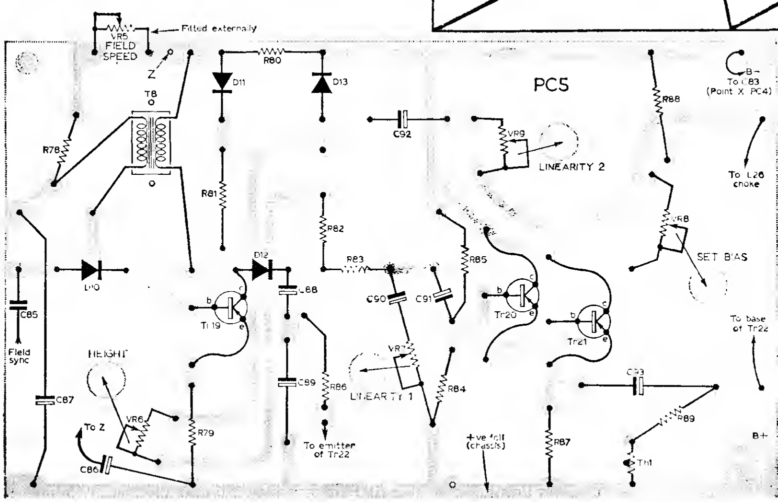


Fig. 36. Correction of the basic field sawtooth for linear scan.

Fig. 37 (below). *The printed board for the field-scan generator (PC5). The copper parts are shown shaded.*



reading with the oscillator and driver unit should be 95–110mA, and when the output transistor is switched in, the meters should read about 0·6A in all. The output transistor should *not* get noticeably warm after 5 minutes. If all seems well the resistor may be shorted out progressively and the ammeter reading should *decrease* so that when completely shorted out the unit takes about 0·4A. The line output transistor, as it operates in a hard-on or hard-off state, should dissipate negligible heat and remain quite cold.

With the receiver obtaining a programme of reasonable field strength, VR11 is set to mid-travel and the output transistor switched on, VR12 should then be set until lock is achieved. If this condition cannot be met, or met only with VR12 near an extremity of its travel, the core of TA should be adjusted. Alternatively, VR12 can be left at about its mid-setting and the core of TA rotated for lock. The "Q" of the tuned circuit is higher with the core more than half-way home, and a larger trigger amplitude is obtained.

Connect an oscilloscope to the test-point, at the junction of R75 and R76, and observe the wave-form of the collector fly-back pulse. With a Perspex rod, gently adjust the position of the two-turn coupling coil (which is wired between the two legs of the e.h.t. transformer) until the third harmonic falls on the peak of the flyback pulse, and the wave-shape of Fig. 23d is obtained. The peak collector voltage is then minimum. If possible, the adjustment should be made with the e.h.t. supplying 100μA load.

It is important to note that the output stage must never be switched on unless the oscillator speed is above 9,000c/s, otherwise the output transistor may be damaged. When making up the receiver as a whole, switching is arranged so that the on-off switch has four positions, as follows:

1—off

2—sound only

3—sound, with line and field oscillators switched on

4—scan output stages brought into circuit.

This allows sound only to be listened to if desired (which means battery saving when using accumulators), and also allows the line-whistle to be heard before the output stage is connected to the supply. This last position of the switch also switches on the c.r.t. heater.

The circuit is a "fail-safe" device. If the oscillator should fail, the base switch-waveform is not generated and the output transistor stays "off".

Adjustment of the e.h.t.

When the Olympic II was in its final development stages and a tube was required, Messrs. Mullard stated that the tube AW36–11 would not be available in this country, but offered the M36–11W (an industrial tube) instead. This policy has now been reversed, and it is understood that the AW36–11 is now available for this receiver. This is in a way a good thing, as

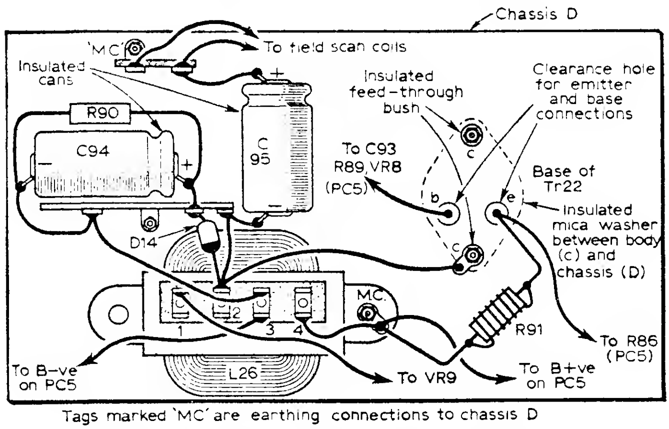


Fig. 38. Layout of the output stage of the field-scan generator. Note that chassis "D" forms the heat-sink for Tr22.

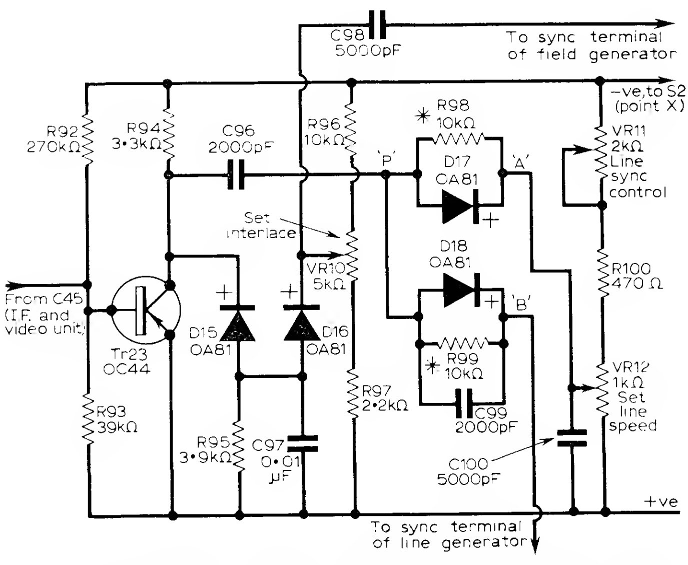


Fig. 39. The circuit of the sync control unit. If desired, R98 and R99 may be reduced to 6·8kΩ with some improvement in sync stability.

the "entertainment" tube is a good deal cheaper than the precision monitor used in the prototype.

However, although the two tubes are very similar electrically, the AW36–11 is rated at 12kV working while the M36–11W takes 14kV. Consequently, if

the receiver is made up precisely to specification there may be some danger of "flash-over" at the higher e.h.t. voltage. While some protection is specified, by way of "decoupling" capacitors from final and focus anodes to the Aquadag coating, a flash-over could damage some components, and is best avoided. An effective way of arranging for the correct e.h.t. potential is to alter the period of fly-back, and this is done by adjusting the value of C79. Although a nominal value of 0·17μF has been specified, the adjustment is made by *reducing* its value from about 0·2μF in steps of about 0·01 to 0·02μF (using parallel capacitors) until the correct safe value of e.h.t. is obtained. This ensures that in the early stages of adjustment *less* than rated e.h.t. is generated, the e.h.t. rising as adjustment proceeds.

A suitable e.h.t. voltmeter is needed, and for those who do not own one a simple and safe design is described in Chapter 23, on page 301.

The line output stage is connected up to the tube in the normal way, and the video (cathode) lead connected to the receiver. C79 is made up of the following capacitors in parallel, all of 250V rating at least: 0·1μF, 0·05μF, 0·02μF (2 of), 0·01μF (2 off).

After switching on, the raster is adjusted to an average value of overall brightness. The timebase generator should be running under synchronised conditions if possible, but if not the speed should be correct or as nearly so as can be judged.

The e.h.t. voltmeter is connected between chassis and the final anode of the c.r.t., and the voltage read off. With the foregoing combination of capacitors, this voltage will probably be of the order of 9kV.

C79 is then adjusted *downwards* in value (switch off between trials) until a value of e.h.t. of between 11kV and 11·5kV is attained, after which the capacitor combination is not changed again. The third harmonic coupling should be re-trimmed for minimum peak collector voltage (oscilloscope across the test-point—i.e. across R76). The adjustment is now complete, and the screening cover may be placed on the line-scan stage.

Although this time base generator does not radiate badly, screening is necessary since pulse rise and fall times are very fast and some energy does escape. Decoupling must be very thorough and if a capacitor is available (and will fit) of more than 100μF, it may well be used, next to the 5·6Ω decoupling resistor (RR) where −12·6V enters the oscillator-driver unit. Up to 1000μF may be used, although this may entail the unit continuing to whistle for a second or two after switching off.

Line Lock

Line lock is extremely stable and very hard. When first set up, the prototype operated for 32 hours, in sessions of up to 2 hours at a time, without any adjustment being required. With a small accumulator as supply, the voltage on load should drop from 12·6 to about 10·5 reasonably steadily, and the sync can cope with this. Further use without recharging results in a

rapid drop to 9V or lower, and sync cannot always be obtained in these circumstances. When set, both line and field controls are semi-permanent and do not require to be readily accessible; this may simplify the layout of the front of the receiver.

If the line hold control is not set too far towards the "out of lock" state, it is not possible to observe by inspection whether a station has ceased to radiate at the end of transmission—this is easy with a normal receiver since sync obviously fails and is both seen and heard. However, the absence of this effect could possibly result in the receiver's being left switched on all night, and is hardly recommended!

Field Scan Circuit

The field timebase generator and output stage also use a blocking oscillator as the driving element but the switching mode is not used in the output device. Study of Fig. 35, the theoretical circuit, will show that the blocking oscillator (Tr19) is used to provide a heavy positive pulse which is fed into a timing circuit consisting essentially of a resistance and capacitance in series. Considering C88, the 4μF timing capacitor as originally discharged (actually it is 4μF [C88] and 25μF [C89] in series) the action is that this begins to charge up to the -12V supply through R82 and VR6 in series. Before this capacitor is fully charged the positive pulse arrives, via D12, the 1S121 diode, and discharges it. Thus a repetitive sawtooth is produced whose amplitude is governed by the extent to which the timing capacitor is charged and whose repetition frequency is governed by the p.r.f. of the blocking oscillator. A further 1S121 diode (D13) is inserted in the charging circuit for reasons to be mentioned later.

Between the sawtooth-generating circuit and the base of the amplifying transistor is interposed a further adjustable network comprising a 2·2kΩ resistor (R83), an 8μF capacitor (C90) and a variable 1kΩ resistor (VR7). The purpose of this is to develop a parabolic wave from the generated sawtooth which, when added to the latter at the base of the amplifying transistor, will pre-distort it so that later on the distorting effect of the output choke will be alleviated. Fig. 36b shows simply how this is done. In fact, further correction is needed to obtain a really linear scan, and the means will be discussed shortly.

The input resistance of transistors operating in the common-emitter configuration is low, and because of this a severe load is placed on the timing capacitor which badly impairs linearity of charging. Consequently a composite transistor is used for amplification. This consists of transistors Tr20 (OC45) and Tr21 (OC81) arranged in a "Darlington pair"; this configuration of two cascaded emitter-followers has an input impedance of the order of a megohm, which imposes negligible load on the timing circuit. From this combination, input for the power-amplifying transistor Tr22, an OC28, is derived.

Darlington Pair

The statement above that the input resistance of the "Darlington pair" in the field scan generator is about 1MΩ refers, of course, to the transistors Tr20 and Tr21 alone; naturally the bias network is in parallel with this, giving about 25kΩ effective input resistance in all.

However, without the Darlington connection the input resistance would be only a few kilohms, if that, and as the charging resistors are small and the value of the timing capacitor large, 25kΩ is relatively negligible as a drain affecting linearity adversely.

As it stands so far, the scan coil current is by no means linear, even with the precautions and correction mentioned above. To improve matters, the initial sawtooth is linearised, so allowing the parabolic correcting wave to exert more effect. For this purpose a second winding is placed upon the output choke, and the voltage developed in this winding is fed back into the timing circuit. The purpose of D13, the second 1S121 diode will now be seen. As the timing capacitor begins to charge, the aiming potential is 12·6V. Because of the amplification arranged the voltage fed back from the choke begins to rise. As soon as the voltage fed back rises above 12·6V the diode cuts off the battery, allowing the choke output to "take over" as the source of aiming potential. Thus as the timing capacitor charges up, the aiming potential rises, and the exponential charging of the capacitor is transformed into a very linear rise.

This obviously can occur effectively only when the diode cuts off the battery supply, and although the main part of the scan is made very linear the first half inch or so (at the top of the picture) still tends to be rather cramped. This is dealt with by means of a further correcting circuit.

Boot-strapping

The feedback described above represents a type of "boot-strapping", and the polarity of the feedback must be arranged appropriately by connecting the feedback windings of the choke (L26) the right way round. This can easily be found by trial and error. The actual amount of voltage feedback is governed by VR9, the 25kΩ potentiometer (linearity 2), and is necessarily in the right phase to maintain oscillation. Complete linearisation by this means involves that the gain in the feedback loop is unity, and this is also the condition for oscillation to occur. Consequently, something short of complete linearity has to be settled for. However, the original pre-distortion of the waveform as generated does not involve feedback, and so complete linearity can be attained providing correct adjustments are made. During the setting-up process, oscillation may readily be obtained but this should cause no undue alarm as the circuit is so arranged that no harm comes of it. However, it should not be allowed to persist very long for obvious reasons!

The slight cramping at the very start of the scan could really be neglected, since plenty of "overscan" is provided for, and if necessary it could be hidden

behind the mask of the tube. However, for best possible display of the Test Card it is worth removing, and this is done by a feedback loop which includes an emitter resistor (R91) of 2Ω in the output transistor circuit and the 25μF capacitor (C89) in the timing circuit. The feedback is controlled by a series resistor of 33Ω (R86) and a moment's consideration of Fig. 36 will reveal that this feedback is operative throughout the scan period. It is most effective of course right at the start of scan, and performs its intended function accurately. This does not need to be made adjustable, but if preferred R86 may be replaced by a 100Ω variable, or 250Ω variable if the former is not readily available.

The best method of synchronisation is via a buffer amplifier, since the "backwash" from the blocking oscillator is thereby prevented as far as possible from affecting the interlace circuit. In order to obtain a pulse of the correct phase, however, a two-stage amplifier would be required, and the cost would not be justified. Accordingly, a diode is used to isolate the blocking oscillator. The range of lock so obtained is entirely adequate, and in practice once set does not need to be altered at all. The speed control (VR5) could therefore be put at the back of the receiver out of the way, if desired.

The current required by the output transistor Tr22 is only 125mA, and this could be supplied by an OC82, OC84 or 2G382. However, type OC28 is specified partly because of its higher collector-to-base voltage rating, but chiefly on account of the fact that it is a much more massive device; the small current taken does not warm it appreciably. Thus, with a good heat sink (the chassis, if one is used) and temperature compensation by means of Th1, the VA1040 thermistor (Mullard) in the base circuit, its characteristics change little when in operation, and negligible change in picture height occurs even after hours of continuous operation. The network comprising an OA81 (D14), 8μF capacitor (C94) and 3·3k resistor (R90) across the output choke L26, serves to minimise the large overswing voltage induced during flyback. A 2kΩ potentiometer (VR8) in the base circuit of Tr22 is for adjustment of bias in setting up. The collector current should be set to 125mA while scanning the tube. A preliminary setting of 120mA may well first be made.

The Printed Circuit

The suggested etched circuit (PC5) is shown in Fig. 37, but there is nothing critical about it and if the constructor can find a simpler layout it might well be used. The circuit is not complete on this board, as with the line scan circuit; the output circuit comprising Tr22, L26 etc. is mounted on a 7in × 4in piece (minimum heat-sink dimensions) of 16 s.w.g. sheet aluminium heat sink as shown in Fig. 38, using the insulating kit provided with the output transistor (which must be ordered with it). The emitter resistor (R91) consists of a length of resistance wire, conveniently wound on a ¼in former and slipped off. The terminal board for L26 can also carry D14 and the voltage linearity CR network R90, C94, but the 1000μF capacitor (C95) is a rather bulky item and a tag strip is best used to anchor it

firmly; this same tag strip can also carry terminals for the scan coil leads. Tr22 etc. may alternatively be mounted on the main chassis, which will then form its heat sink.

Setting-up for rough linearity can be accomplished on a received picture, or by applying an oscilloscope (Y-amplifier input) across a 1Ω or 2Ω resistor in series with the scan coils. Near-perfect linearity can be achieved only on the Test Card, and is done by trial and error. It will be found best to reduce height a little, so that a 1in gap appears at top and bottom of the picture. For this preliminary setting, VR9 should be set to a point well below that at which oscillation sets in. Best linearity should then be sought by means of VR7 (linearity 1), the 1kΩ potentiometer in the correcting circuit. Then VR9 should be adjusted, followed by adjustment of the height control VR6, to the correct value. Small further adjustments of both variable linearity controls will achieve a display of excellent linearity and at the correct height.

It is quite important to see that the OC28 collector current is adjusted to 125mA before the above adjustments are finalised.

The prototype employs as a blocking oscillator transformer a Mullard ferrite assembly type LA1. It was handy, small in size, and gives accurate timing in the blocking oscillator circuit. However, this is an obsolescent component, and its up-to-date alternative is not cheap. Hence an alternative specification is given in Table E, which has been tried in the circuit and found satisfactory.

The thermistor Th1, used in series with R88 (330Ω) in the base circuit of Tr22 is not always stocked by local dealers, but can be obtained without much delay at low cost. This is for correction for ambient temperature, and for testing purposes the series combinations may be replaced by a 470Ω resistor. If this is left in circuit, however, it will be found necessary to alter the timebase height and linearity controls from time to time, as the room temperature changes.

It is emphasised that nothing in this receiver gets hot—except of course the c.r.t. heater and the heater of the DY86. The transistor which runs the hottest is in fact the video amplifier Tr8, and this gets barely warm to the fingers in long use. As the total receiver consumption is only 9W or so, the above will be understood. Heat sinks are provided for the transistors carrying fair currents, but these are not to prevent destruction of the transistors concerned but to ensure steady junction temperatures as far as possible.

Setting-up the Field Generator

In the first place, when construction has begun, it is as well to complete building the field oscillator before any other circuits are added to the etched board. Then a check should be made to see that the transformer connections have been made the right way round. In winding, primary and secondary will normally be in the same direction, and if this is the case the "outermost" wire of the primary goes to the collector and the "outermost" wire of the

TABLE E

Blocking Oscillator Transformer T8 (as in Prototype): Core: Mullard type LA1 assembly Windings: Secondary (put on first) 600 turns 42 s.w.g. enamelled copper wire, random wound. Primary, 1200 turns same wire, random wound. Interleaving—1 turn Sellotape *Blocking Oscillator Transformer T8 (alternative specification):* Core: 0·5in stack of No. 18 E & I Silcor 1 laminations Windings: Secondary (put on first) 540 turns 42 s.w.g. enamelled copper wire, random wound. Interleaving—0·002in	paper, two layers. Primary, 1100 turns 42 s.w.g. enamelled copper wire, random wound *Output Coke L26:* Core: $\frac{3}{4}$in stack No. 35 E & I laminations Silcor 2 (M & EA) Gap: 0·002in Windings: Choke winding: 450 turns 26 s.w.g. enamelled copper wire Feedback winding: 600 turns 40 s.w.g. enamelled copper wire Interleaving 2 turns Sellotape *2-ohm Resistor R91:* $6\frac{1}{2}$in length of 34 s.w.g. Eureka wire

secondary to the base of the transistor. The two innermost ends of the windings (the centre tap, if the windings had been continuous) are to the negative supply and the speed-control resistor respectively. If correctly connected in circuit, the stage will be heard to oscillate if a pair of headphones are wired across R81, the 820Ω resistor in the collector circuit, or other suitable pick-up point. When this is correct, the remainder of the generator may be built.

The remaining units required for the completion of the receiver are the sync control unit, the flyback blanking unit and the linearity control of line scan. The first includes the device responsible for sync separation, the production of field (frame) sync pulses, and a phase-connecting line sync circuit which gives a d.c. output suitable for controlling the line oscillator. The second comprises the network which provides the requisite direct voltage for the first anode of the cathode-ray tube, an adjustable voltage for the focus anode, and brightness control.

The flyback blanking unit hardly deserves the name of "unit" as it comprises a few resistors and capacitors which feed a pulse to the c.r.t. grid to blank out the flyback lines which might otherwise be seen on the tube face.

Sync Control Unit

The theoretical circuit is shown in Fig. 39. Tr22 (OC44) is operated with a base bias supply which sets it at its correct working point. The transistor cuts off when the video signal arrives at the base, while passing current only as the sync pulse train occurs in the opposite polarity to the video signal. Across R94 (3·3kΩ) the collector load resistor there appear just the sync pulses, at considerable voltage. The field sync pulses are separated by a conventional "interlace diode"—in fact two diodes D15 and D16 (OA81) contribute to the function, as seen. The bias on this pair of diodes is so

arranged as to cut off the line sync pulses but to pass the much larger field sync pulse which is produced by the action of the "time-constant" circuit R95 (3·9kΩ) and C97 (0·01μF). From this circuit the field sync pulses pass, via C98 5nF, to the sync-input terminal of the field generator.

The line sync pulses are derived also from the video signal and appear in a positive-going sense, limited in amplitude, at the collector of Tr23 (OC44). These are rectified across circuit "A" in the diagram, and the bias across D17 (OA81) determines the amplitude of the resulting steady voltage.

However, a constant-amplitude negative-going pulse from the line output transistor, shaped and attenuated by a CR network, is also rectified across circuit "B" and contributes to the voltage developed at point P. If these two pulses coincide they behave as one pulse, since their amplitude is constant, but if they occur at different times the total pulse width, and hence the voltage developed at point P, increases. This increases the total base current into the base of TrA (line scan generator) and the speed of the generator tends to rise until the pulses coincide, when the action ceases. If the speed of the line scan generator decreases, therefore, the action of the circuits described is to speed it up to synchronisation point. Should the speed of the generator tend to increase from synchronisation, nothing will stop it rising until the pulses are quite distinct and a constant voltage is generated. No harm is done by this, however; the synchronisation just fails. It is therefore necessary to arrange that at all times the tendency is for the speed of the line scan circuit to be less than that of the sync pulses. This is similar to conventional practice with valve circuits, and the arrangement behaves in a very similar way.

The suggested etched circuit is shown in Fig. 40. It will be noted that the output from "B" circuit—the line sync connection—is not a true sync output in itself; it serves the dual purpose of carrying d.c. sync output to the first transistor of the line scan circuit and of transferring a shaped pulse from the scan unit to the sync unit.

To set up the unit, it should be connected to the vision i.f. amplifier and the line scan generator, and the assembly switched on, but NOT the line output transistor. The method of receiving the line pulse harmonics, as radiated from the line oscillator timing circuit, on a synchronised domestic TV may be used to keep a check on line scan speed. Otherwise the speed may be judged by ear, listening for the magneto-striction whistle from the blocking oscillator and driver transformer.

VR11 (line sync control) is set at about the centre of its travel, and VR12 (set line speed) is then adjusted to obtain the correct oscillator frequency, adjusting the core of TA so that lock is obtained at the mid-position of VR12.

If an oscilloscope is available, it should be connected between the field sync terminal and "earth", and the field sync pulses displayed. When VR10 (set interlace) is correctly set, Fig. 41 is the display obtained—the line sync pulses are just cut off. If noise is appreciable, it will be well to adjust the control so that the sync pulses are appreciably shorter—this cuts off some of the noise.

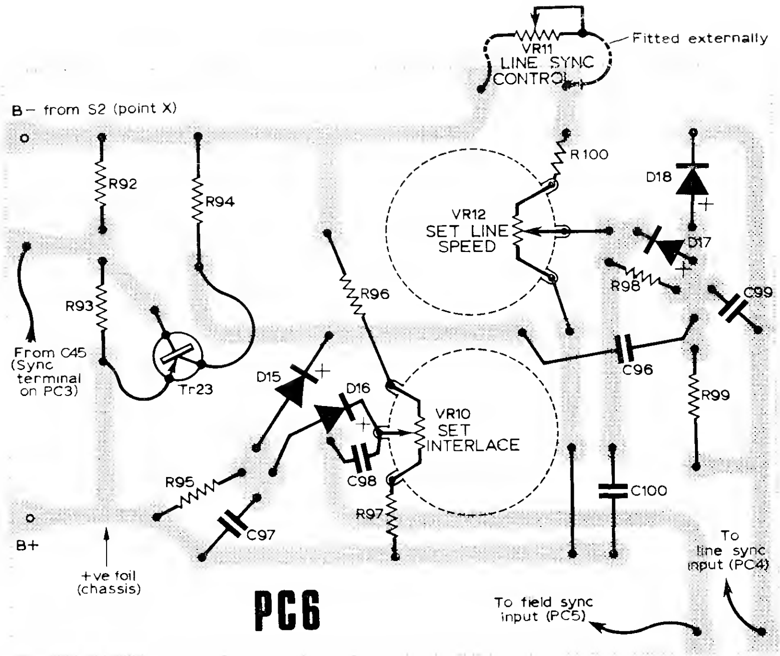


Fig. 40. The printed board of the sync control unit (PC6). The copper parts are shown shaded.

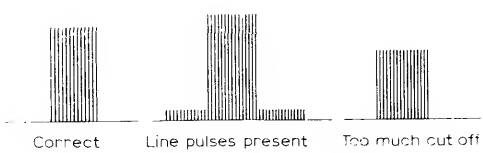
If an oscilloscope is not available, the "set interlace" control VR10 should be set to its mid-point temporarily, and later—when a picture is being displayed—adjusted for best interlace. In noisy circumstances, advantage will be had by reducing the degree of field lock available, by rotating the 5kΩ potentiometer VR10 until the field speed control has a suitable range of movement.

It will be found that both line and field "hold" controls are quite "hard", and that no difficulty is experienced in obtaining a stable well-interlaced picture.

The c.r.t. Control Unit

This small unit is strictly not essential, for the various elements could well

Fig. 41. Display of field-sync pulses.



be dispersed among the other units already built. However, it has the advantage as a separate circuit that all the supplies to the c.r.t., except heater and e.h.t., are controlled at one point and this is very handy for servicing and layout. It is not expected that any major servicing will be needed, for the chief cause of failures is usually heat in one place or another and this receiver develops negligible heat. However, volume controls and other variable resistors do wear out in time!

The theoretical circuit is shown in Fig. 42, and the suggested etched circuit in Fig. 43.

There is nothing special about the etched circuit layout, which may be altered as desired as long as high-voltage points are well insulated. Note that there is no battery-negative connection. The socket which is mounted on one shorter end of the circuit board is one of the small and popular non-reversible plug-and-socket components widely available for connecting h.t. and l.t. batteries to radio receivers. Only the socket is here used, to enable a a suitable focus voltage to be selected. The range of focus voltage available is -75V to $+500$V and the tap giving best results should be selected for the particular tube supplied. In the prototype, $+500$V gives the finest focus, but other tubes of the same type may need another setting. A small cut-out is made in the etched board to accommodate the socket. The wander plug to fit the sockets is taken from the plug supplied, soldered on to the end of a flex lead long enough to stretch from its printed-circuit-board connection to any of the sockets A, B, C or D.

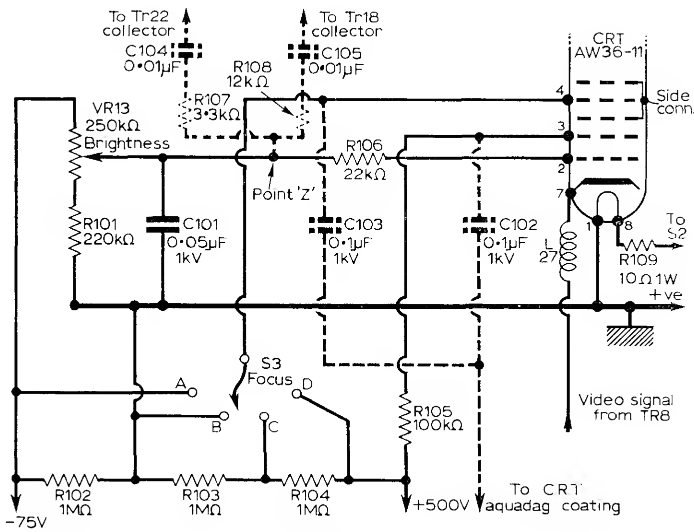


Fig. 42. Circuit of the c.r.t. control unit (the parts shown dotted are not on PC7).

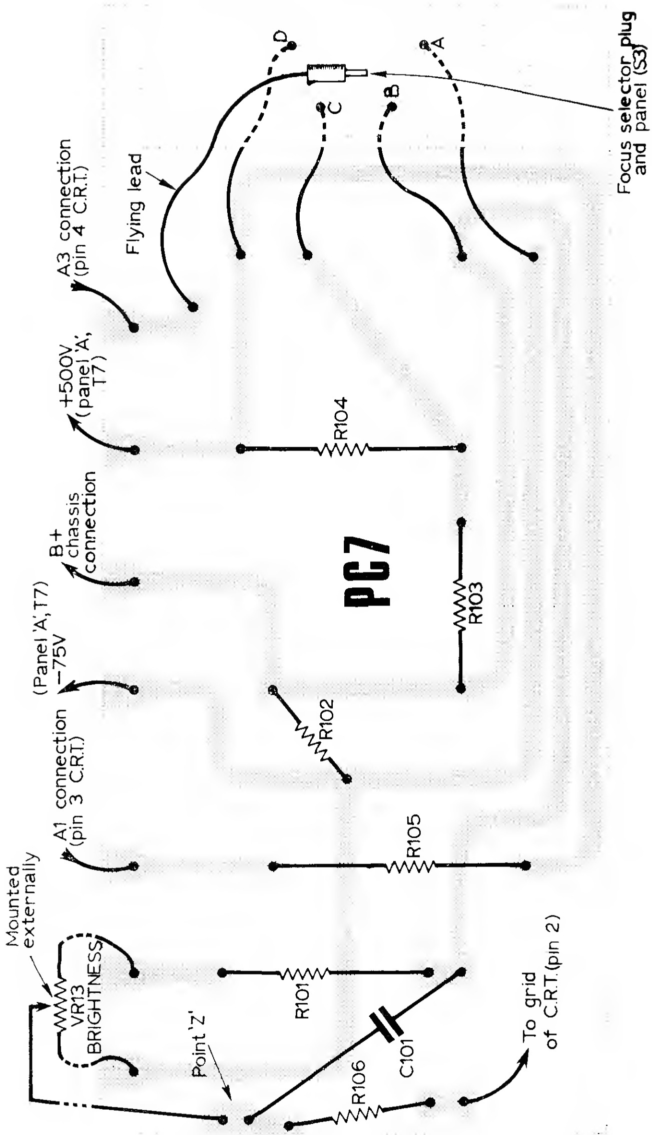


Fig. 43. The printed board of the c.r.t. control unit (PC7). The copper parts are shown shaded.

This small unit is, in the prototype, mounted below the chassis on 4B.A. screwed rod. It may, however, be put in any convenient place, according to receiver layout.

The Flyback-Blanking Unit

Fig. 44 shows the way the c.r.t. grid connection is wired to receive the flyback pulses from the two scan generators. This assists in removing from the picture the return traces of the scanning sweeps. There are, of course, blanking pulses in the transmitted wave-form, and these reach the cathode of the c.r.t. However, their amplitude is not great in any case, and since the video amplifier is called upon to give a relatively high voltage output for video it may well be worth working on a non-linear part of its characteristic as far as these pulses are concerned. Hence, their amplitude tends to be less than necessary unless such an aid is provided.

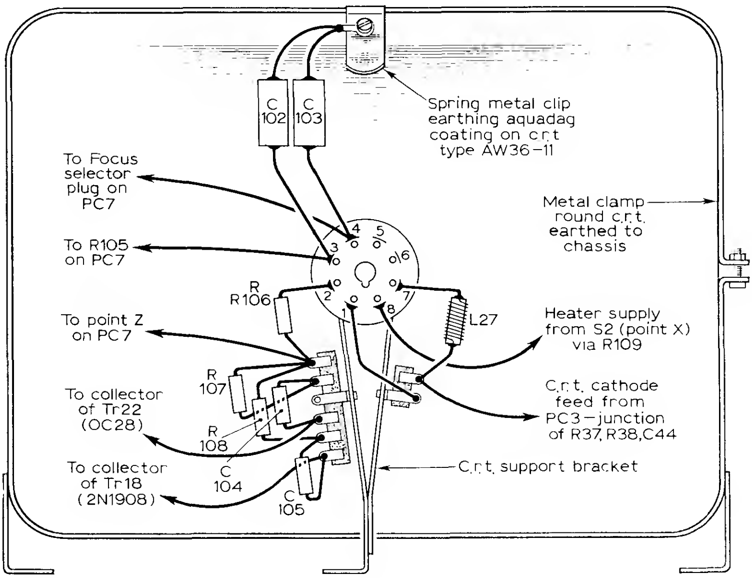


Fig. 44. The base connections of the c.r.t. and the layout of the flyback-blanking circuit.

Line Scan Linearity Control (L28)

Because of resistances in the line output stage—either direct ohmic resistance or by way of losses—the early part of the scan is not quite linear enough to give a perfect picture. Accordingly, a conventional correcting circuit is employed which makes use of a saturable reactance. The required components are Mullard core type FX1054 and magnet M2529. These may be

subject to delay in delivery, and in the prototype a 1in length of $\frac{1}{4}$in diameter ferrite aerial rod was used instead, together with a couple of small bar magnets. This worked quite well and is still in use. L28 comprises 48 turns of 20 s.w.g. enamelled copper wire, wound in two layers each held firmly with Sellotape. It is wired in series with the deflector coils. The adjustment required is to magnetise the core sufficiently, that almost as soon as scan current flows it saturates and its inductance drops to quite a small value. This is done by mounting one bar magnet (or perhaps two, if temporary arrangements have to be made!) close to the reactor. The polarity needed is found by trial and error, and so is the position of the magnet. Once adjusted, the two can be bound together with Sellotape; no further alteration will be required. The connections to the scan coils and L28 are given in Fig. 45.

Assembling the Receiver

Certain precautions are necessary when the various units of the receiver are collected together into the relatively small compass dictated by practical considerations. Room has to be found for the loudspeaker, in addition to the assemblies already described, and while the location of the units is not critical, there are certain things which must not be done, and others which should be done.

To begin with, the line output transformer must not be too close to the middle part of the cathode ray tube. There is a considerable magnetic field associated with the transformer, and it varies from the beginning to the end of the scan. Although it has a "closed" magnetic circuit, there is a good deal of leakage in practice, and distortion of the raster is possible. The best place on a limited chassis is as far away from the tube as convenient, and nearer the base of the c.r.t. than elsewhere. If some distortion is encountered a small bar magnet can correct it quite well. Its position must be found empirically.

It is also important that the line scan transformers are not too near the a.f. transformers of the sound receiver but much more is it necessary to separate the latter from the field scan components. If too close, a pronounced hum at 50c/s will be heard in the loudspeaker.

The tuner unit can go nearly anywhere, and so can the sync unit and the c.r.t. control unit. The various adjustments needed can be made on the bench, and there will be no call for alteration later.

Because of the effects of ambient temperature, it is preferable to mount the "hold" controls on the front of the receiver. In the prototype they are fitted with smaller knobs than the main controls, to distinguish at a glance.

The function switch is arranged as in Fig. 46. The reason for this is two-fold. In the first place, it may well happen that only the sound is required, especially when using an accumulator with the receiver. The second "on" position enables the line oscillator to attain its steady frequency before the output stage is switched on. It might well be possible to do without this

intermediate switching, but the writer has not cared to risk it! The line whistle can be heard very faintly, so this switch position ensures that the operator knows all is well before switching in the line output transistor.

It is worth remarking that this receiver uses a 14in cathode ray tube requiring 90° deflection, and therefore does not come into the category of receiver using a tube requiring only 42° deflection, for which the line scan requirements are quite modest. While such tubes have their place, this receiver requires the output transistor to operate close to its limits, and although a good safety margin has been left it will be prudent at all times to exercise care.

Fig. 47 shows the c.r.t. neck support bracket.

Power Supplies

Since either accumulator or car battery can provide power, as well as the mains unit already described, a chassis-mounted plug should be provided. The free sockets, on the end of appropriate leads, can then be plugged-on, whichever supply it is desired to use. There is difficulty in obtaining such components commercially, and the usual non-reversible plugs and sockets used for domestic electricity are not suitable because of the serious chance that some day somebody will plug the receiver directly into the mains. The solution to this problem is one which constructors will have to solve out of their own resources. There are some reversible components on sale, but these are best avoided for obvious reasons. If use has to be made of such components, a power diode *MUST* be wired in series with the supply so that if a plug and socket are connected the wrong way round no harm will be done. For this purpose, the collector-base junction of a germanium power transistor is very suitable. If a silicon rectifier is used, the IS411 is suitable; its voltage drop on load is not high and little effect will be noticed on the working of the receiver. The chassis can act as the heat sink for this diode, and it will then be best to wire it in series with the positive-supply lead.

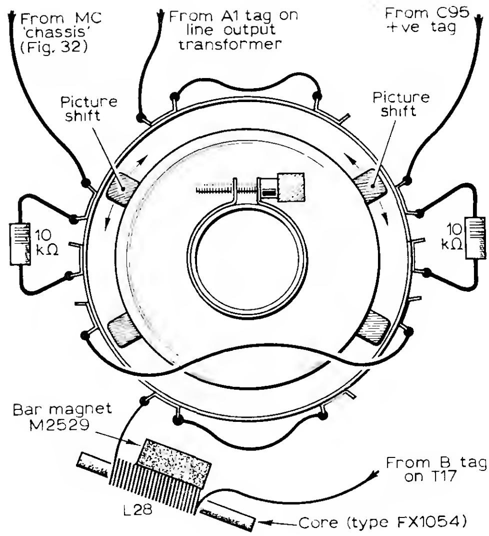
A suitable non-reversible three-pin plug is made by Bulgin and can be obtained from the usual sources of supply.

Aerials

Obviously a good aerial is an advantage, but for portable use a telescopic dipole has proved quite satisfactory in field trials. Owing to the exceptionally low noise figure for the receiver, quite rudimentary aerials have been found to give good results, but it is worth mentioning that the aerial bandwidth must be sufficient if resolution is not to be impaired.

For channels 1 to 5, $\frac{1}{4}$in tubing is recommended; heavy flex separated out may well do at a pinch but it is unlikely to give better than 1·5–2Mc/s resolution. For B and III frequencies, however, flex can be quite satisfactory.

Wherever possible the aerial should be no nearer the receiver than 4ft.



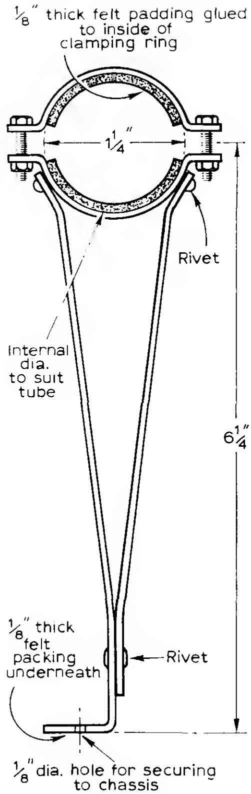
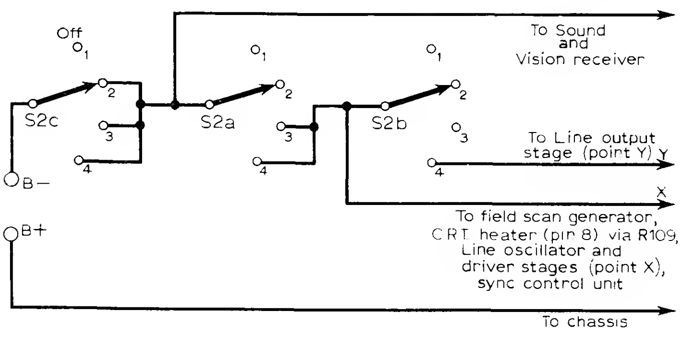


Fig. 45 (above). *The connections of the scan coils.*

Fig. 46 (below). *Circuit of the function-switch S2.*

Fig. 47 (right). *Supporting-bracket for the neck of the c.r.t.*



In spite of good screening some of the "radiated" energy from T7 may escape, and may well result in the appearance of faint vertical lines on the raster. Actually, this field is inductive rather than radiative, and falls off in intensity as the cube of the distance of the aerial from the receiver, so an extra few inches may make all the difference.

Final Adjustments

If the receiver units have been aligned as and when they have been constructed, little remains to do when the whole assembly is complete. Doubtless sound only will have been received for some time before the time comes to check overall operation, so that will be one worry out of the way.

The moment of truth arrives, however, when the picture is to be switched on as well, and the likelihood is that on turning up the brilliance control a good picture will be seen at once, needing only the adjustment of the line linearity control and the video correcting choke L22. It should cause no dismay if mirror-imaged or upside down, since the connections to the scan coils can be reversed without difficulty.

If, however, the image is broken up and negative in appearance, the vision detector diode has been joined into circuit the wrong way round; this should, of course, have been detected in preliminary setting-up operations, and is not likely.

The components should conform to specification, where one is given. For example, if AF116 transistors were used in the i.f. stages instead of the specified OC171, virtually a complete re-design would be involved, especially in the neutralising circuits, and the overall gain would be very different from what was expected.

Specifications have therefore not been given unless necessary, for one 100μF 12V working electrolytic capacitor will not differ much in its electrical characteristics from one of another make. However, good quality components are most desirable, lest failure should have dire consequences in damage to other components. Mostly the design is "fail-safe". A good deal of attention has been paid to this point, but there are of course positions in the circuit where it cannot be done except at great expense or inconvenience, as for example in the case of the brilliance control potentiometer. If during construction equal care is given to soldering and the placing of components, a sound electrical and mechanical job will result whose performance will not disappoint. It is confidently claimed that this receiver is in a class of its own as regards picture and sound quality, consistency of performance and versatility of application.

Automatic Gain Control

The receiver which has been described in the preceding pages was not provided with a.g.c. since this presented a major design problem if it were to be done properly. However, after the completion of the prototype, an a.g.c. system was designed and full details of the extra circuitry and its design now follow.

The provision of automatic gain control for a television receiver is undoubtedly a refinement which might well be optional. In the first place, unless the signal strength is quite high there is always a tendency towards a "grainy" if not "noisy" picture, because of inherent receiver noise on the

one hand and interference on the other. As a result—unless one is a DX enthusiast—the favourite reception is of stations giving a high field strength, and these are necessarily near enough to be not normally subject to severe fading.

Aircraft Flutter

Nevertheless, any programme can be sadly marred by the effect of "aircraft flutter", a form of interference that has always been with television and is likely to get worse in the future. This phenomenon is due to wave-interference arising from double-path reception, and is often accompanied by the formation of positive or negative "ghosts" on the picture. Few effects, except perhaps severe mains-borne static are more unpleasant, since in many commercial receivers (in fact, in the majority) such consequences as rapid variation in picture size as well as brightness and contrast appear, together with break-up of the picture, vision-on-sound and sound-on-vision due to an overloaded input stage, and so on. Automatic gain control, in an area where aircraft are a nuisance, is thus a definite advantage.

Automatic gain control in the vision channel can be had for next to nothing if one is not too fussy about it, but really effective a.g.c. is not quite so cheap. The favourite method of control in commercial practice is "mean-level" control, and this can be had (for example) at the grid of the sync separator valve. The trouble with this is that the mean level of the signal is by no means the ideal reference, since it necessarily varies with the content of the picture. Thus, a picture containing large dark masses has little modulation on it, and if the control-voltage depends on mean level the gain of the receiver overall will increase—producing a rather noisy raster together with a largely grey screen. It all comes right again when a more highly modulated signal reappears, the correct black-level being restored approximately. However, a picture whose noise and black-level depend on picture content can hardly be regarded as satisfactory. An audio analogy would be a.g.c. which operated on modulation rather than on carrier; speech and music would be reduced to a monotonous level whose chief variation would be in noise.

Sync Pulses

The only part of the transmitted waveform which is maintained at constant level for a given field strength is the sync-pulse modulation; the amplitude of the sync pulses at the detector does vary with the aerial signal and nothing else. This is therefore the only part of the received signal which will give correctly arranged a.g.c.

There are two ways in which the sync pulses can be made use of. In the first place, line sync pulses occur regularly, and during the pulses there is no picture modulation. Hence, if line sync pulses can be sampled for amplitude, an error signal can be obtained which is independent of picture content.

Secondly, during the field-sync period, field sync pulses only appear. Again, if these pulses are sampled for amplitude, an error signal independent of the picture can be had. In principle, either sampling can be done readily enough by means of suitable circuitry. The a.g.c. unit developed for the Olympic II is of the field-sync sampling type, and the reason for this is as follows.

To sample the amplitude of a line sync pulse, a "gating" pulse has to be provided which effectively connects an amplifier to the video signal during the sync pulse. The gating pulse must therefore occur after the "front porch" has begun and must finish not later than the end of the "back porch", and can thus be only about 18μs in width. It recurs every 98·7μs. Any delay in starting the pulse will lose some of the 18μs, and any delay in terminating the pulse will mean that some of the picture content of the next line will be included in the "sample". Delays in transistorised circuits tend to be rather large because of the effects of changing base charge, and although the thing can be done it is not simple. The likeliest way would be to use the line sync to gate an amplifier to sample the sync pulse, but as soon as an amplifier is used to get enough of a pulse to do the gating, the delays reduce the available pulse width, and the system is insensitive.

Longer Time

However, the field sync pulses last for a much longer time, and an amplifier delay of a few microseconds wastes very little of them. This is economical, relatively easy to get going, and cheap. The circuit diagram is given in Fig. 48, and the method of operation should be relatively easily understood from the following brief description.

The amplifying transistor is Tr25 whose function is to deal with the video signal. It is arranged to operate in the common-emitter mode; R114 and R113 are the base-bias resistors and R115 provides emitter bias in the usual way. Between the base of the transistor and R113 a further transistor Tr24 is interposed, and this is normally cut off because its base is returned to the negative rail. The transistor is an *npn* type for convenience in circuit arrangement.

Since Tr24 is cut off, Tr25 base receives no d.c. bias and is also cut off. However, the base of Tr24 is connected through the pulse-shaping network R110/C106 to the collector of the field output transistor, and when the field scan is completed, the flyback pulse causes the base of Tr24 to draw current—the transistor is in fact switched hard on by the pulse. This action connects d.c. bias to the base of Tr25, and Tr25 cuts on.

Video-signal Gating

A suitable network R111/C107 connects the base of Tr25 to the emitter-follower section of the video amplifier, and when Tr25 cuts on, the whole of the video signal is applied to the base of Tr25. An amplified version of

the signal then appears at the collector of Tr25, and is rectified by the diodes D19 and D20, and smoothed by C110, R118, and C111. A voltage proportional to the video signal is thus developed across VR14. Since during the field-flyback pulse the video signal consists of sync pulses varying between approximately black level and zero, the object of the exercise has been achieved.

The voltage developed across VR14 is a negative-going voltage because the diodes D19 and D20 are appropriately wired into circuit. Part or all of this voltage is applied to the base of a further transistor Tr26, and base current flows. The effective resistance between collector and emitter of Tr26 is dependent on the base current, and thus as far as direct current is concerned, Tr26 becomes a resistor whose value decreases with increase of signal strength.

Modifications

Referring now to Fig. 13, page 67, VR1 may be removed and the junction of R20 and C23 connected to the collector of Tr26. At the same time, for best results, R19 should be increased in value to 18kΩ. R110 is connected direct to the "live" (non-earthy) end of the field-output-choke feedback winding—see Fig. 37, where the lead may be connected to the circuit-board PC5 at the edge-connector labelled "To L26 choke". Video signal to the base of Tr25 is supplied by soldering a 4·7kΩ resistor (R111) to the circuit-board shown in Fig. 14 (PC3) at the point labelled "Sync", with a lead from the resistor to C107 in Fig. 48.

The suggested printed circuit for the a.g.c. unit is shown in Fig. 49. VR14 may be secured to the board if desired in the position shown, or it may be arranged at the end of not-too-long leads. VR14 is a "set-a.g.c."

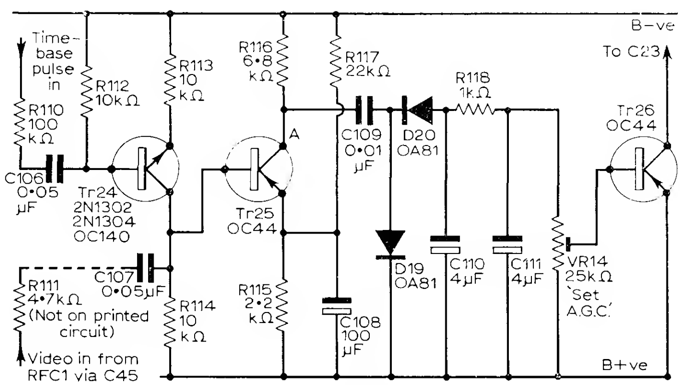


Fig. 48. Circuit of the a.g.c. unit.

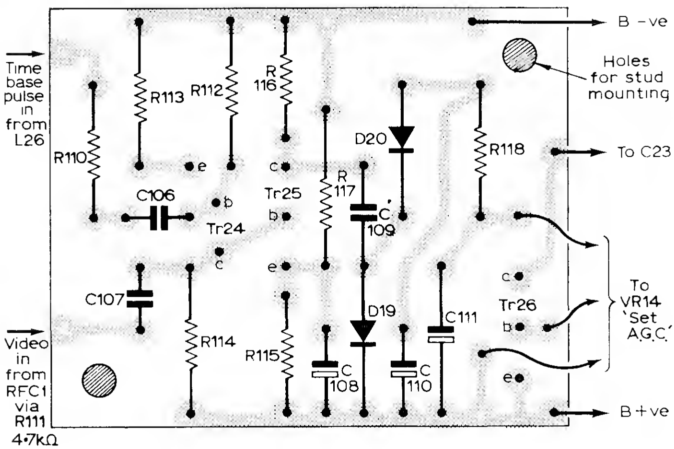


Fig. 49. The printed board of the a.g.c. unit.

control, and is intended to be pre-set, but, if required, may be adjustable by the user of the receiver.

Control of Contrast

It is important to realise that VR14 cannot be used as a contrast control, since it is inside the a.g.c. feedback loop. It can be correctly adjusted only when a further modification to the receiver has been carried out; since by taking out VR1 the set has been deprived of its contrast control, it is now necessary to put in again such a control, this time outside the a.g.c. feedback loop.

This is readily done by removing R34 (Fig. 13) and substituting a 2kΩ variable resistor (VR15) in series with a 33Ω resistor (R119), as shown in Fig. 50. The slider of the variable resistor is connected to "earth" via a 100μF capacitor (C112) and this variable resistor now constitutes the contrast control. Also, R36 is changed from 12kΩ to 6·8kΩ.

VR2 (15Ω—see Fig. 13) is next readjusted so that the transistor Tr8 takes 2·5mA collector current.

Control of contrast is effected by varying the negative feedback applied across VR15; part or all of this is decoupled by C112, and so the gain of the output stage can be varied by rotating the slider without altering the direct current flowing. Control is effective from an overall gain of 3 to an overall gain of 25 for the stage.

Setting VR14

To set up VR14 in Fig. 48, the contrast control is set for maximum gain and a milliammeter placed in series with R18 (Fig. 13). VR14 is adjusted

until a current of 1·5mA is obtained, with a good signal being received at the aerial. No further adjustment is required. Changes of signal strength of 25dB (voltage) at the aerial input produce changes of video output of 6dB (voltage), and thus a control range of about 10:1 is achieved. This, over a single i.f. stage, is adequate for most purposes.

If an oscilloscope is available, it is instructive to use it to ascertain the waveform at the point A in Fig. 48 (the collector of Tr25). The appearance of the display should be closely similar to that in Fig. 51, all trace of picture-content waveform being removed between field-sync pulses, while during field pulses the field-sync waveform appears.

It is important to see that no spurious pulses appear at the collector of Tr26. While these would not appear on the picture (since they occur during field flyback) they might have an effect on the d.c. level at the collector of Tr26 and so affect the a.g.c. action. C110 and C111, together with R118 and VR14, represent a small time-constant which will deal with most aircraft flutter quite well. If however C110 and C111 had to be increased to remove spurious pulses, the time-constant would be lengthened and the a.g.c. made slower in action. In such a case, it would be best to connect a further $4\mu F$ capacitor between the slider of VR14 and "earth", setting the slider some way from its position of maximum travel. This would certainly lengthen the time-constant, but by adding virtually another filter element would do so in the most effective way and with minimum increase in time-constant.

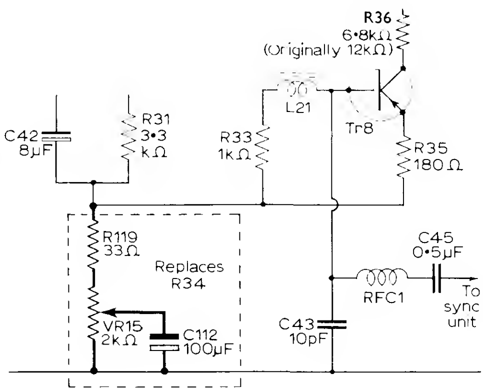


Fig. 50. Alterations to the circuit of Fig. 13 to accommodate the a.g.c. circuit.

Fig. 51. The waveform at the collector of Tr25.

625-Line

It should be realised that what follows effectively states the problems in converting the receiver to 625-line operation: how they will be solved is another matter.

There are three important matters to consider in the 625-line conversion: namely, the increased receiver video bandwidth to be achieved, with appropriate gain; the fact that frequency-modulated sound is the means of transmitting audio information; and lastly the increased number of picture lines and the effect of this on the working of the line-output transistor. These will be discussed in turn.

The r.f. tuner need cause no concern since the bandwidth for each channel will be adequate for the reception of 625-line transmissions, should these ever be undertaken on any of the Channels 1 to 13. Operation at u.h.f. will require a separate u.h.f. tuner, which will probably be bought from a commercial source.

As to the i.f. and output sections, these will require some modification unless it is desired to construct new printed circuit boards which can be switched in or out as required. This may be simpler, though more expensive, and it will be for the constructor to decide whether he wishes to convert or duplicate.

To ease design problems, which can be time-consuming, it has been arranged that the increased bandwidth needed in the vision receiver can be obtained by altering the inter-stage coupling. Referring to Fig. 13, C26 and C31 will be altered in value, and C31 attached to the tap (4) on L17. The sound rejector system L18, C32, C33, R26 will be removed, and C36 may need to be altered in value.

The damping resistor Rx may also need to be reduced in value, and the polarity of the diode D1 reversed. This will involve an alteration in the mode of operation of the video amplifying stages Tr7 and Tr8.

It is probable also that inter-carrier sound will be the best way to arrange audio reception, but initially this will not be attempted, since the twin superhet already exists and will only need altering in frequency. However, a discriminator will have to be provided instead of the detector D2.

Line Output

The line output stage is, of course, a quite serious problem, although rather simpler of solution than the vision receiver. The calculations resulting in the design of the scanning yoke and e.h.t. transformer are tedious and will not be quoted here, but briefly the results are that with 405-line transmissions the peak collector voltage is about -65V under the conditions specified, while peak collector current is of the order of 7A.

With 625-line working the peak current is about the same, but the peak collector voltage rises to about 100V, and this results in a state of affairs in

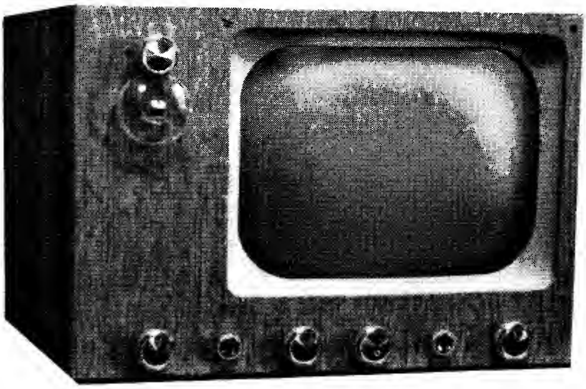


Fig. 52. The finished Olympic II transistorised TV.

which the output transistor is operating nearer its limit of 130V. This is safe, of course.

However, the scan period is much reduced and in order for peak currents to reach the required value the e.h.t. transformer must be used as an output transformer in the same way as is done with valve circuits. This involves a tap on the e.h.t. transformer primary; it is not difficult to modify the specified transformer in this way, but currents are heavy and switching from 405 to 625 standards will involve the use of a very low-resistance switch if linearity is to be retained. Also, the third-harmonic tuning conditions will be changed by the modification, and extra coupling turns will have to be switched in.

It is emphasised that although all the foregoing may seem complicated, the design problems are quite capable of solution. However, some experimental work is going to be needed as well as calculations, and readers are not advised to "have a go" in advance unless they are very much aware of the details of the physical situation!

Note: See page 120 *et seq.* for Specification and Lists of Components.

The Olympic II Transistorised TV Receiver

Specification

Power Supplies:

12V d.c. (accumulator for normal use with motor-car or in caravans, etc., or dry battery Ever Ready type TV1 in emergency)

190–250V a.c. (separate power unit for a.c. mains, for use as domestic TV). Interchangeable polarised power socket provides for quick change from domestic to mobile use

No adjustment of transformer needed for a.c. mains unless supply voltage drops below 190V

Power Unit:

(For use with a.c. mains) fully stabilised

Supply variable from 11V to 16V. Hum level less than 20mV

Current available 0–3A

Power Consumption:

9W on 405-line operation. On 625-line operation, consumption will be approximately 25W

Standards:

To receive 405-line transmissions on Bands I and III

Capable of conversion to 625-line standards at relatively low cost at a later date

Display:

14in picture tube, Mullard type A36–11W (electrostatic focus, electromagnetic deflection both line and frame, 90° deflection)

Tuner Unit

Six channel switch, with fine tuner (electronic or capacitive)

3-transistor circuit

I.F. outputs—Sound 38·15Mc/s
—Vision 34·65—37·65Mc/s

Noise figure 5dB or better

Sound Receiver:

Two i.f. stages, diode detection and noise limiting, audio pre-amplifier, driver and Class AB push-pull output

Audio output 1 watt (conservative estimate)

Distortion (total harmonic) 3% at 1W

Loudspeaker 3Ω

Overall sensitivity 3μV at aerial socket for 50mW output

Vision Receiver:

Three i.f. stages, diode detection, 2-stage video amplifier

Overall sensitivity 10μV at aerial socket for 20V p-p at c.r.t. cathode

Bandwidth 3Mc/s

Sound-channel rejector (main)—infinite-rejection bridged 1-1T circuit

Time Base Generators:

Field: Blocking oscillator, driver, output, with linearising circuits; choke coupled to scanning yoke. Direct locking. Linearity better than 8%

Line: Blocking oscillator, driver, output switch (transistor). Direct coupled to scanning yoke, part of transistor current diverted to e.h.t. transformer. "Flywheel" sync. Linearity better than 5%

Internally Generated Power Supplies:

−75V for video amplifier
+500V for tube supplies
+14kV e.h.t.

Assembly:

The prototype is in unit construction, for reasons of development. However, an integrated assembly, using the chassis for all heat sinks would save weight and space. Unit construction is, however, very flexible and permits of individual adaptation as required; it will also permit of more simple conversion to 625-line standards at a later date. Etched circuits are used widely in the prototype

List of Components

Tuner

Resistors:

R1 1kΩ
R2 1·2kΩ
R3 2·2kΩ
R4 10kΩ
R5 270Ω
R6 3·3kΩ
R7 330Ω
R8 1·5kΩ
R9 1·2kΩ
R10 8·2kΩ
R11 470Ω
R12 8·2kΩ
R13 1·2kΩ
R14 1·2kΩ
R15 220Ω
R16 330Ω
All 10% ¼W carbon

Capacitors:

C1 1500pF tubular ceramic
C2 70pF silver mica
C3 1500pF tubular ceramic
C4 1500pF tubular ceramic
C5 1500pF tubular ceramic
C6 1500pF tubular ceramic
C7 3·9pF silver mica
C8 4·7pF silver mica
C9 8·2pF silver mica
C10 1500pF tubular ceramic
C11 1500pF tubular ceramic
C12 3·9pF silver mica
C13 1500pF tubular ceramic
C14 1500pF tubular ceramic
C15 1·5pF tubular ceramic
C16 1000pF ceramic feed-through
C17 1·5pF tubular ceramic
C18 1500pF tubular ceramic
TC1 1·5–5pF tubular ceramic trimmer
VC1 1–5pF air spaced variable fine tuning control, see text and Fig. 3.6 for alternative design)
All fixed capacitors ±20%

Transistors:

Tr1 *Texas 2G102 or Philco 2N1742
Tr2 Texas 2G102 or Philco 2N1743
Tr3 Texas 2G102 or Philco 2N1743 or Mullard AF102
Mullard AF102 may be used in all three stages, with a slight reduction in gain

Switches:

S1a/b/c 3-pole 6-way (on two 2-pole 6-way wafers, see text). Radiospares "Makaswitch" assembly

Inductors:

L1–L2–L5–L6, L9–L10–L11 } Band I coils
L3–L4–L7–L8, L12–L13–L14 } Band III coils
T1 Band I transformer
T2 Band III transformer
1FT1 1st i.f. transformer (tuned to 37·5Mc/s)

Miscellaneous:

PC1–PC2: Printed circuit board (copper clad laminate)—Bakelite Limited type DH74 or similar, two pieces, one is 3in × 2in the other 3in × 3in
0·3in dia polystyrene former (L9)—1
0·3in dia bakelite former 2¼in in length (T1–T2)—2
Can assembly ¾ × ¾ × 1⅜in, with 0·3in former (i.f.t. 1)—1
V.H.F. iron dust slugs (purple-coded)—4 (to fit L9, T1 and i.f.t. 1)
Brass slugs (0B.A.)—2 (T2)
Quantity of nuts and bolts and washers, 6B.A. brass; connecting wire etc.; 22 s.w.g. aluminium sheet (for the screening box); perforated zinc, pk screws

* *Texas Instruments Ltd., Manton Lane, Bedford*

LIST OF COMPONENTS

I.F., VIDEO AND AUDIO STAGES

Resistors:

R17 3·3kΩ
R18 100Ω
R19 10kΩ
R20 470Ω
R21 1kΩ
R22 3·3kΩ
R23 10kΩ
R24 100Ω
R25 1kΩ
R26 2·7kΩ
R27 100Ω
R28 10kΩ
R29 3·3kΩ
R30 1kΩ
R31 3·3kΩ
R32 4·7kΩ
R33 1kΩ
R34 470Ω
R35 180Ω
R36 12kΩ
R37 220kΩ
R38 100kΩ
R39 2·2kΩ
R40 10kΩ
R41 1kΩ
R42 100Ω
R43 10kΩ
R44 2·2kΩ
R45 1kΩ
R46 100Ω
R47 33Ω
R48 8·2kΩ
R49 4·7kΩ
R50 10kΩ
R51 4·7kΩ
R52 39kΩ
R53 1·8kΩ
R54 10kΩ
R55 1·5kΩ
R56 82kΩ
R57 470kΩ
R58 10Ω
R59 680Ω
R60 3·3kΩ
R61 33Ω
*R62 3Ω
R63 150Ω

All 10% ¼W carbon

* Three 10Ω in parallel

Potentiometers:

VR1 5kΩ wire-wound (contrast control)
VR2 15kΩ miniature carbon pre-set
VR3 10kΩ carbon (volume control)

Transistors:

Tr4 OC171
Tr5 OC171
Tr6 OC171
Tr7 OC171
Tr8 AF118
Tr9 OC171
Tr10 OC171
Tr11 OC75
Tr12 OC81D } LFH3 package
Tr13 OC81 }
Tr14 OC81 }

Diodes:

D1 OA70
D2 OA70
D3 OA70

Miscellaneous:

T3 Driver transformer Radiospares T/T6
T4 Output transformer Radiospares T/T7
PC3 Printed circuit board (copper clad laminate) 12in × 6in
Long can assembly (0·3in dia. former) 2
Short can assembly (0·3in dia. former) 7
V.H.F. iron dust slugs (purple) ... 11
Standard iron dust slugs 4
0·3in dia. polystrene formers ... 4
Quantity of 6B.A. nuts, bolts and washers. Connecting wire. Two co-axial sockets, etc.

Capacitors:

C19 0·01μF paper
C20 180pF silver mica
C21 1500pF tubular ceramic
C22 5pF silver mica
C23 0·01μF paper
C24 8pF silver mica
C25 1500pF tubular ceramic
C26 100pF silver mica

List of Components

I.F., Video and Audio Stages (*cont.*)

Capacitors (*cont.*)

C27 5pF silver mica
C28 1500pF tubular ceramic
C29 1500pF tubular ceramic
C30 1500pF tubular ceramic
C31 1500pF tubular ceramic
C32 120pF silver mica
C33 120pF silver mica
C34 8pF silver mica
C35 1500pF tubular ceramic
C36 100pF silver mica
C37 18pF silver mica
C38 1500pF tubular ceramic
C39 220pF silver mica
C40 5pF silver mica
C41 5pF silver mica
C42 8μF electrolytic 15V
C43 10pF silver mica
C44 1μF paper
C45 0·5μF paper
C46 5pF silver mica
C47 47pF silver mica
C48 470pF silver mica
C49 1500pF tubular ceramic
C50 1500pF tubular ceramic
C51 25pF silver mica
C52 25pF silver mica
C53 1500pF tubular ceramic
C54 1500pF tubular ceramic
C55 25pF silver mica
C56 25pF silver mica
C57 1500pF tubular ceramic
C58 33pF silver mica
C59 8μF electrolytic 15V
C60 47pF silver mica
C61 8μF electrolytic 15V
C62 0·01μF paper
C63 100μF electrolytic 15V
C64 100μF electrolytic 15V
C65 16μF electrolytic 15V
C66 1000μF electrolytic 15V
C67 100μF electrolytic 15V
C68 0·2μF paper

Variable Capacitors:

TC2 2–10pF air spaced trimmer
TC3 2–10pF air spaced trimmer

List of Components

Power Pack

Resistors:

R1 220kΩ
R2 330Ω
R3 1·5kΩ
R4 3·9kΩ
R5 330Ω
R6 330Ω
R7 470Ω

Capacitors:

C1a 2000μF elec. 25V
C1b 2000μF elec. 25V
C2 4μF elec. 15V

Miscellaneous:

Tr1 OC35 (Mullard) or any power transistor of 6A collector rating or higher (requires mica insulating washers)
Tr2, Tr3 OC81 (Mullard) or 2G382 (Texas)
D1, 2, 3, 4 1S411 (Texas) (D2 & D4 require mica insulating washers
D5 1S2068 (Texas)
T1 200/250V primary, 16V 2A secondary
Neon, miniature type (GEC)
F1, 2 1A cartridge fuses

List of Components

Line Timebase

Resistors:

RA	390Ω
RB	22Ω
RC	100kΩ
RD	4·7kΩ
RE	1kΩ
RF	1kΩ
RG	5·6kΩ
RH	470Ω
RI	100Ω
RJ	22Ω
RK	4·7kΩ
RL	220kΩ
RM	39Ω
RN	100Ω
RO	4·7kΩ
RP	33Ω
RQ	0·5Ω*
RR	5·6Ω
R64	2·2kΩ
R65	390Ω
R75	10kΩ
R76	100Ω
R77	10kΩ

All 10% ½W

* No. 34 s.w.g. bare Eureka, 1½in (coiled as necessary)

Note: The following resistor-numbers are not used: R66, R67, R68, R69, R70, R71, R72, R73, R74 and VR4

Capacitors:

CA	100μF 15V electrolytic
CB	2μF 15V electrolytic
CC	0·1μF paper
CD	0·5μF paper
CE	0·1μF paper
CF	100μF 15V electrolytic
CG	100μF 15V electrolytic
CH	5nF paper
CI	100μF 15V electrolytic
CJ	0·1μF paper
CK	0·1μF paper
CL	0·5μF paper
C69	0·01μF paper
C70	0·05μF paper
C71	10μF 15V electrolytic
C77	0·5μF paper
C78	0·1μF paper
C79	0·17μF 350V paper 5%
C80	0·1μF 750V paper
C81	8μF 150V electrolytic
C82	100μF 15V electrolytic
C83	330pF silver mica 10%

Note: The following capacitor-numbers are not used: C72, C73, C74, C75 and C76

Transistors:

TrA	OC44
TrB	2G382 or 2G381 or OC84
TrC	2G382 or 2G381 or OC84
TrD	2G382 or OC84
TrE	2S301 or 2G382 or OC84
TrF	2G382 or OC84
Tr18	2N1908 with insulating tape

Note: The transistor-numbers Tr15, Tr16 and Tr17 are not used

Diodes:

D4	OA81
D5	OA81
D6	2SO2P
D7	1S411
D8	BY100
D9	OA202

Transformers:

TA Core, Denco type 9A, primary 350 turns, secondary 75 turns, both 42 s.w.g. enam. copper wire

TB Blocking oscillator transformer (see text), 2-FX1238 cores required—Mullard Ltd.

TC Driver transformer (see text), 2-FX1239 cores required—Mullard Ltd.

T7 E.H.T. transformer LOP/4188 Elac Ltd.

Note: The inductor-number L25 is not used. Also, transformer-numbers T5 and T6 are not used

Miscellaneous:

Scanning coils: Line coils 140μH, Frame coils 37Ω, Elac Ltd.

V1 DY86 Mullard Ltd.

Aluminium sheet, 16 s.w.g. 7in×4in (chassis "C"), 7in×4¾in (metal panel "B"). 18 s.w.g. two pieces 2½in×2in for heat sinks. PC4—copper clad laminate 4in×3¾in. Polythene sheet, perforated zinc sheet, transistor mounting clips—2. Quantity of 4B.A. bolts, nuts, solder tags and washers, 4B.A. studding, ⅜in coil former fitted standard dust core (L25), quantity of 24, 26 and 32 s.w.g. enamelled copper wire (for L25, T5 and T6 windings). 1½in length of 34 s.w.g. bare Eureka wire (for R74)

List of Components

Field Scan Generator

Resistors:

R78	47kΩ	R86	33Ω
R79	4·7Ω	R87	220Ω
R80	33Ω	R88	2·2kΩ
R81	820Ω	R89	330Ω
R82	1·5kΩ	R90	3·3kΩ
R83	2·2kΩ	R91*	2Ω
R84	39kΩ	All 10% ½W	
R85	68kΩ	* See text	

Potentiometers:

VR5 50kΩ carbon
VR6 2kΩ wire-wound
VR7 1kΩ wire-wound
VR8 2kΩ wire-wound
VR9 25kΩ carbon

Capacitors:

C85 0·05μF paper
C86 500μF electrolytic 15V
C87 2μF electrolytic 15V
C88 4μF electrolytic 15V
C89 25μF electrolytic 15V
C90 8μF electrolytic 15V
C91 16μF electrolytic 15V
C92 100μF electrolytic 15V
C93 500μF electrolytic 15V
C94 8μF electrolytic 150V
C95 1000μF electrolytic 25V

Transistors:

Tr19 OC44
Tr20 OC45
Tr21 OC81
Tr22 OC28
(with insulating kit)

Diodes:

D10	OA81	D12	1S121
D11	OA81	D13	1S121
		D14	OA81

Miscellaneous:

Th1 VA1040 thermistor—Mullard Ltd.
T8 Blocking oscillator transformer (see Table 5)
L26 Field output choke (see Table 5)
Aluminium sheet 16 s.w.g. 7in × 4in (chassis "D")
PC5 Copper clad laminate 7½in × 4½in
Quantity of 4B.A. nuts, bolts, solder tags and washers, tag-strips, quantity of 26, 40 and 42 s.w.g. enam. copper wire (for L26 and T8). 6½in length of 34 s.w.g. Eureka resistance wire (R91)

List of Components

Sync-control Unit

Resistors:

R92	270kΩ	R97	2·2kΩ
R93	39kΩ	R98	10kΩ*
R94	3·3kΩ	R99	10kΩ*
R95	3·9kΩ	R100	470Ω
R96	10kΩ	All 10% ½W	

Capacitors:

C96 2000pF mica or paper
C97 0·01μF paper
C98 5000pF mica or paper
C99 2000pF mica or paper
C100 5000pF mica or paper

Transistors and Diodes:

Tr23 OC44
D15, 16, 17, 18 OA81

Potentiometers:

VR10 5kΩ carbon
VR11 2kΩ wire-wound
VR12 1kΩ wire-wound
PC6 Copper clad laminate 4½in × 3½in
* If desired, R98 and R99 can be reduced to 6·8kΩ with some improvement in sync stability.

List of Components

C.R.T. Control Unit

Resistors:
R101 220kΩ
R102 1MΩ
R103 1MΩ
R104 1MΩ
R105 100kΩ
R106 22kΩ
R109 10Ω

Capacitors:
C101 0·05μF 1kV
C102 0·1μF 1kV } Fitted at c.r.t. base
C103 0·1μF 1kV } (not on PC7)

Inductor:
L27 60 turns of 40 s.w.g. enamelled copper wire-wound on a ¼in diameter former, with no core

Miscellaneous:
VR13 250kΩ carbon
S3 4-pin socket with plug
PC7 Copper clad laminate 4in × 1⅝in
c.r.t. Mullard AW36-11 (B8H base)

List of Components

Flyback-Blanking Unit

Resistors:
R107 3·3kΩ
R108 12kΩ
All 10% ½W

Capacitors:
C104 0·01μF paper
C105 0·01μF paper
1-6-way tag-strip

List of Components

The A.G.C. Unit

Resistors:
R19 Originally 10kΩ, now 18kΩ
R36 Originally 12kΩ, now 6·8kΩ
R110 100kΩ
R111 4·7kΩ
R112 10kΩ
R113 10kΩ
R114 10kΩ
R115 2·2kΩ
R116 6·8kΩ
R117 22kΩ
R118 1kΩ
R119 33Ω
All 10% ½W
VR14 25kΩ
VR15 2kΩ

Capacitors:
C106 0·05μF paper
C107 0·05μF paper
C108 100μF 15V electrolytic
C109 0·01μF paper
C110 4μF 15V electrolytic
C111 4μF 15V electrolytic
C112 100μF 15V electrolytic

Transistors:
Tr24 2N1302, 2N1304, OC140
Tr25 OC44
Tr26 OC44

Diodes:
D19 OA81
D20 OA81

Section 3

AERIALS

Chapter Seven

MAKING TELEVISION AERIALS

THE function of a receiving aerial for television is much the same as that of the aerial used for reception of sound broadcasts on the domestic radio receiver; namely, to collect a strong enough signal to operate the set satisfactorily. Of course, the situation is somewhat different on the medium and long waves commonly used for broadcasting; signals travel large distances and are still easily picked up by a simple aerial such as a short length of wire. However, on the short wavelengths, or high frequencies, used for television, the signals do not travel so far and consequently more efficient aerials are required for television than for ordinary broadcast reception. As the frequency employed is raised, the range of the transmitter tends towards what is known as the "optical". The meaning of this expression is clear if one imagines that the transmitter is replaced by a very powerful light: all the area illuminated receives signals, the brightest parts, or areas of good reception being nearest the transmitter. Those parts in shadow, behind a hill, mountain, or other large object, are out of sight of the transmitter, and there, reception is either impossible or extremely difficult.

Signal Strength

Many different kinds of aerial are employed for television reception and which type is used in a particular location depends largely upon the strength of the signal received there. In general, the stronger the signal, the simpler the aerial. However, it is not possible to prophesy which type of aerial will be the best one for a given location as so much depends on local conditions. This fact is often illustrated; sometimes an indoor aerial suffices in one house while across the road a five-element array is required to secure an acceptable picture. Thus, the selection of the aerial is possible only by a combination of experience, trial and error. This fact is rarely appreciated and should therefore be noted carefully.

For first experiments, at least, therefore, the aerial design selected must be such that the aerial can be modified to give better results if necessary. A design which fulfils this requirement is the Yagi aerial, named after a Japanese engineer who carried out much of the early work on this type of aerial. Yagi aerials are often termed "arrays" and vary in complexity from the two-element to those using 14, 16 or even more elements.

Television aerials are tuned devices made up of inductance and capacitance just like any electrical tuned circuit. With an ordinary tuned circuit made

up of a coil of inductance L and capacitance C an increase in tuned frequency is accomplished either by taking turns from the coil (to decrease the value of L) or by reducing the value of C (or both).

When the tuned frequency is decreased, of course, L or C (or both) is made larger.

Frequency and Wavelength

The frequency is the number of times that a complete cycle of waveform occurs in one second. The wavelength of the signal is the distance taken by one complete cycle (see Fig. 1).

Radio waves, in common with the other types of so-called electro-magnetic waves such as light rays, X-rays and cosmic rays, travel (in free space) at the speed of 186,000 miles per second. The metric system is usually adopted and the equivalent of 186,000 miles is 300,000,000 metres. We can thus say that radio waves travel at a speed of 300,000,000 metres per second.

Thus, the wavelength is equal to the distance that the wave travels in one second divided by the number of times that the wave occurs in one second. Wavelength is measured in metres or centimetres, and is usually denoted by the Greek letter lambda (λ).

The relationship between frequency and wavelength can be expressed as a formula:

$$\lambda=\frac{300{,}000{,}000}{f}$$

where λ is in metres, and f is in cycles per second (c/s).

Alternatively,

$$\lambda=\frac{300}{f}$$

where λ is again in metres, and f is in megacycles per second (Mc/s).

As a practical example, the wavelength corresponding to a frequency of 50Mc/s, in Band I, is equal to 300/50, or 6m. At 200Mc/s, in Band III, the wavelength is 1·5m; at 500Mc/s in Band IV, it is 0·6m; and at the very top of Band V, it is 30cm.

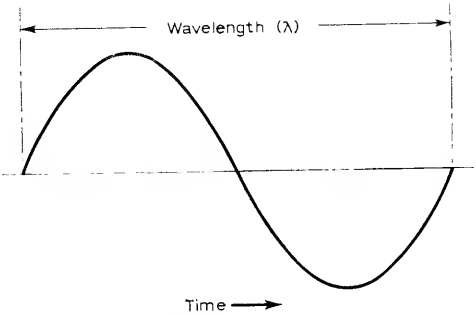
To complete this exercise, we can note that the frequency of a signal of a given wavelength can be found from:

$$f=\frac{300}{\lambda}$$

where f is in Mc/s and λ is in m.

Now if we have a rod that is equal in length to the wavelength of the signal we wish to receive we obtain very good conditions for inducing into the rod almost the maximum possible signal strength. The effect is that a "standing wave" would appear across the rod from one end to the other as shown in Fig. 2.

This wave can represent the change of signal current in the rod and it will be seen that the current starts at zero, rises to a maximum positive value



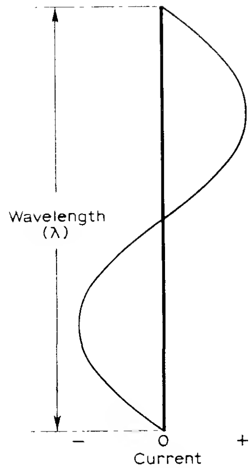


Fig. 1 (above). *The wavelength of an electromagnetic wave is the distance between one whole waveform, as shown. The frequency is the number of times that the wave repeats in one second.*

Fig. 2 (right). *A standing-wave of signal induced into a rod approximately equal in length to the wave.*

and then drops to zero again. This scheme is repeated but this time the current rises to a maximum negative value. We thus have two half-waves which, apart from polarity, are identical from the current-change aspect.

Half-wave Dipole

From a practical point of view, this means that if we have a rod which is only a half wavelength long the signal pick-up efficiency is about the same as that of a rod a full wavelength long.

This is just as well as otherwise there would be twice as much ironmongery adorning the roofs of houses all over the world as there is today! As a practical example the rods of, say, a Channel 1 aerial would all approximate 21ft 4in in length instead of the typical 10ft 7in for that channel.

We have seen, therefore, that to increase frequency (reduce wavelength) on an aerial we have to reduce its length. This action effectively reduces the L and C elements which are possessed by aerials.

It follows that if we have an aerial tuned to, say, 50Mc/s an aerial tuned to 25Mc/s would be double the length and one tuned to 100Mc/s would be half the length. 50Mc/s is in Band I and 200Mc/s is in Band III, which means that a Band III aerial is approximately a quarter of the length (depending upon the exact channel) of a Band I aerial. 800Mc/s is in Band V, which means that a Band V aerial is approximately a quarter of the length (again depending upon the exact channel) of a Band III aerial.

Very approximately, therefore, we can say that the u.h.f. aeiials for BBC-2 have rods which are about an eighth of the length of those on the BBC-1 aerials and about a quarter of the length of those on the ITV-1 aerial. To give *very approximate* figures, a Band I aerial may be 8ft long, a Band III

aerial 2ft long and a Band V aerial 1ft long. These lengths should not be used for constructing aerials as they are given simply to illustrate the approximate ratios of rod lengths over the three bands.

Velocity Factors

When the wave of a radio signal comes up against a tuned aerial, the wave tends to slow down a little from its free-space velocity of 300,000,000 metres per second. This also happens when the signal from an aerial is directed along a feeder cable for application to the receiver. The aerial and cable are said to have a "velocity factor" which is less than unity which corresponds to the unrestricted free-air propagation.

We need not bother too much about velocity factors and the like; it is sufficient to know that the reduction in wave velocity means that for optimum tuning the length of the aerial rod does not quite correspond to the length of the wave. The aerial rod is, in fact, generally made about 95% of the length of the signal wave.

Other factors are involved here; for example the aerial insulator, the manner in which the aerial rod is connected to the feeder and the proximity of the aerial to objects such as other aerials, the roof, metal pipes and so forth. All these things tend to have some influence on the ideal length of the aerial rod and hence the tuning of the aerial. The length of any aerial then is based on some sort of compromise.

A signal wave induces into a rod whose length approximates that of half the length of the wave, both voltage and current waveforms as shown in Fig. 3. This diagram represents a typical half-wave aerial from which it will be seen that the voltage is at a maximum at each end of the rod while the

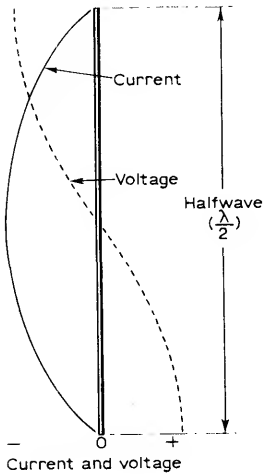
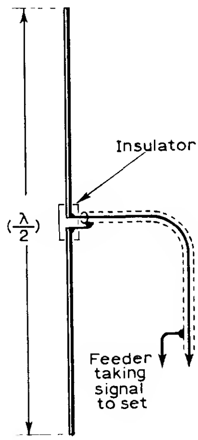


Fig. 3 (left). *A signal wave induces into a rod whose length approximates to that of half the length of the wave both current and voltage waveforms, as shown.*

Fig. 4 (right). *A half-wave dipole comprises two quarter-wave rods separated at the centre by about a quarter of an inch.*



current is at a maximum at the centre of the rod. Should the length of the rod differ substantially from the signal half-wavelength then these conditions will no longer hold. This is rather an important point to keep in mind.

Centre

The most popular arrangement for extracting signal from such an aerial is to sever the rod in the middle, at the minimum-voltage/maximum-current point, and connect the break across a suitable feeder cable that will convey the aerial signal to the set without too much being lost on the way.

We thus have two rods each approximately a quarter of a wavelength long which are supported mechanically at the centre by an insulator. The insulator holds the ends of the two rods about a quarter of an inch apart and is made of a material that does not short-circuit the signal. The arrangement is usually referred to as a "half-wave dipole".

Impedance

Since a dipole is concerned with both current and voltage, it possesses impedance, the value of which is equal to the voltage-to-current ratio. At the centre of the dipole the impedance must be low, since the current is high and the voltage low. If the voltage were zero then the impedance would be zero. In practice, however, there is always a little voltage, even at the centre, and this, in conjunction with the current at the centre, gives an impedance around 72Ω. The impedance rises, of course, up the rods and at the end it is high—several thousand ohms.

While some aerials are fed from the ends (end-fed), most TV aerials are fed at the centre (Fig. 4).

With this in mind, coaxial cable (and twin feeder) is made to suit the 72Ω centre impedance of a dipole, that is, the cable itself has a characteristic impedance of 72Ω. A cable of impedance to match that of the aerial is essential to avoid signal losses and reflections which could impair the reception, especially of BBC-2.

If the impedance of the dipole at the feeder connecting-point differs substantially from that of the feeder cable then a transformer must be connected between the dipole and the feeder to restore the matching or else some other impedance-matching artifice must be adopted, as we shall see.

Polarisation

One more factor needs to be taken into account; the polarisation of the waves from the transmitter. This governs whether the receiving aerial needs to be placed with the plane of the elements vertical or horizontal. The quickest way of determining the correct mounting is to inspect the aerials on your neighbours' houses to see whether they are horizontal or vertical.

Aerials are often said to be "directional". This means that, when mounted, they have to be moved or rotated until the signal received is as strong as

possible. Generally, they are then pointing towards the station (see Fig. 5) but, sometimes, a signal is reflected from an obstacle and this reflected signal is stronger than the direct signal and then the aerial is pointed to receive the reflected signal (see Fig. 6). It should be noted that a simple, *vertical*, dipole is not directional and receives signals equally well from any direction. However, a *horizontal* dipole is directional and must be arranged for best reception, generally broadside on to the direction of the station (see Fig. 7). It must be stressed that arrays do not necessarily have to be pointed direct at the station; much depends on local conditions and the only method is to move the aerial to point in different directions until best results are obtained.

Type of Array

It is often stated that to secure good results in fringe areas or areas of weak signal strength, it is necessary to use the most complicated array that can be devised. This is rarely the best policy; as the array becomes more complicated, its length increases and approaches a wavelength, reflection effects become more troublesome and better results may be obtained with a simpler array. In difficult situations, moving the array a few feet, or raising it slightly, will often enable a good picture to be obtained.

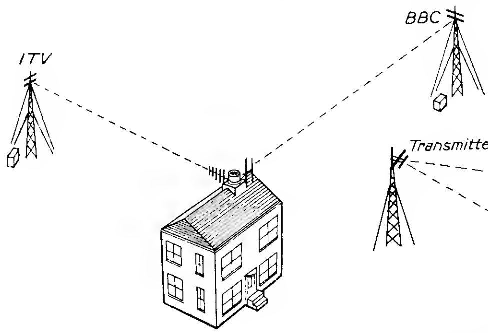


Fig. 5 (left). *Yagi aerials are directional and must be pointed towards the transmitter.*

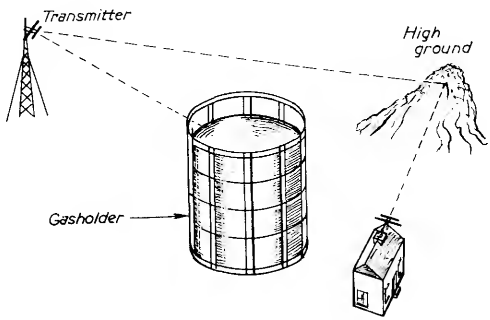


Fig. 6 (above). *If there is an obstacle between the transmitter and the receiving aerial, a reflected signal can sometimes be utilised.*

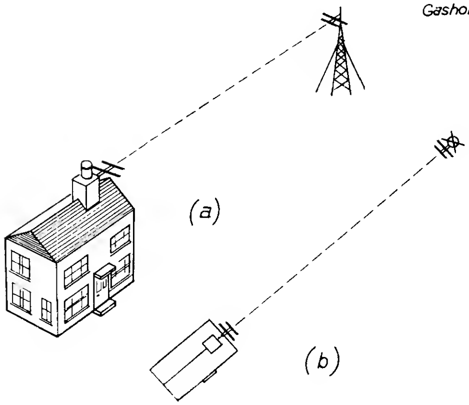


Fig. 7 (left). *Unlike vertical dipoles, horizontal dipoles are directional and need to be positioned broadside-on to the transmitter.*

It should now be clear that in areas known for their poor reception, and in other districts too, the best way of obtaining good results is trial and error. Although the difficulties may seem formidable in the light of what has been said above, in the majority of instances there is no difficulty in obtaining good results at the first attempt.

AERIALS FOR BANDS I AND III

Construction

The materials used in the construction of the aerial depend on the location of the aerial; near the transmitter, the crudest materials—wire and wood—suffice, while where signal strength is low, and every precaution must be taken to prevent loss of strength, better quality materials must be employed. Copper-plated rod can often be purchased cheaply and the "real thing"—aluminium or duralumin rod or tube—may be used for ambitious arrays. If the array is to be used outside, then, of course, construction is more difficult as the insulation needs to be good and the array must weather any storms etc. On the whole, it is advisable for the majority of amateurs only to construct indoor aerials as, without extensive facilities, the construction of sound, outside aerials is extremely difficult, tedious and often disappointing.

Figure 8 illustrates various types of Yagi aerial with their dimensions indicated by letters. Table A gives the dimensions required for reception of the various BBC-1 and ITV stations. It will be noted that no details are given of BBC aerials with more elements than three. Such aerials would be very large and impossible to mount or accommodate.

Directors and Reactors

The illustrations in Fig. 8 proceed from a simple dipole, through the "H" aerial to complicated multi-element arrays. The element added to form the H aerial is called a reflector and, broadly speaking, it increases the signal from the aerial. In each case, the transmitter is assumed to be to the right of the array and the reflector is seen always to be the element furthest from the transmitter. The elements progressively added in front of the dipole are called directors and increasing their number increases the signal picked up and makes the array more directional. Adding more reflectors, however, is found to make little improvement and thus only one is used, even in the nine-element array.

Insulation

The various elements do not need to be insulated from the supporting cross-bar, provided they are mounted or bolted at their mid-points. The cross-bar can be made of wood or metal depending on the weathering qualities required.

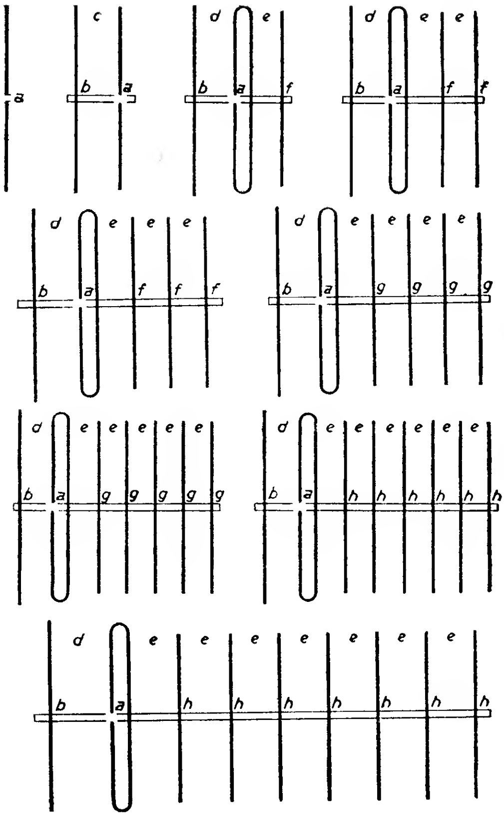


Fig. 8. Various aerials ranging from a simple dipole to a 9-element Yagi. The letters "a" to "h" refer to the lengths of the elements or to their spacings. As it is not practicable to construct multi-element aerials for Band I channels, dimensions "g" and "h" are not given in the Table below.

TABLE A

DIMENSIONS OF THE AERIALS ILLUSTRATED IN FIG. 8

Channel	a		b		c		d		e		f		g		h	
	ft	in	ft	in	ft	in	ft	in	ft	in	ft	in	ft	in	ft	in
1	10	11	11	7	5	8	4	7	3	5	10	5	—		—	
2	9	4	10	0	4	9	3	11	2	11	9	0	—		—	
3	8	6	9	1	4	6	3	7	2	8	8	2	—		—	
4	7	9½	8	4	4	1	3	3	2	5½	7	6	—		—	
5	7	2½	7	8	3	9½	3	0	2	3	6	11	—		—	
8	2	6	2	8	1	3½	1	0½		9½	2	4½	2	4	2	3½
9	2	5	2	7	1	3	1	0		9¼	2	4	2	3½	2	3
10	2	4½	2	6	1	3	1	0		9	2	3	2	3	2	2
11	2	3½	2	5½	1	2½		11½		8¾	2	2½	2	2	2	1½
12	2	3	2	4½	1	2		11½		8½	2	2	2	1½	2	1
13	2	2½	2	4	1	2		11		8¼	2	1½	2	1	2	0½

From the information given, it should be possible to construct an aerial suitable for most locations and the procedure is to choose the design from Fig. 8 and then to determine the dimensions from Table A. The dimensions have been calculated to give good results; it is very unlikely that they will be exactly the same as those employed by commercial aerial manufacturers or those mentioned in other publications. This fact should be borne in mind if any comparisons of element lengths and spacings are made.

AERIALS FOR BANDS IV AND V

In the London area, BBC-2 has a vision frequency of 567·25Mc/s and a sound frequency of 573·25Mc/s, corresponding to Channel 33. This represents a mean frequency, between the sound and vision carriers, of 570·25Mc/s.

It is often desirable to design an aerial for mid-channel working, particularly in service areas. In fringe areas it may be best to design more towards the vision carrier frequency. However, it should be remembered that a u.h.f. aerial is often designed to work over a number of channels. This will be considered later.

Let us suppose that we want a dipole tuned to the mean frequency of Channel 33. The wavelength in metres is equal to 300/570·25; that is, 0·526m, or 52·6cm. To convert 52·6cm to inches, we divide by 2·54. This works out to 20·7in, or 1ft 8·7in. For a half wavelength, we divide by two, which gives 10·3in. A practical aerial will have a length which is 95% of this value, so we multiply by 0·95, which gives an overall dipole length of 9·8in.

Actually, there are one or two approximations in the above working, but at least it serves clearly to reveal the basic principles involved in the calculations.

The whole process may be simplified and the length of the dipole calculated from:

$$l=\frac{468}{f}$$

where l is the overall length of the dipole, in feet, and f is the frequency in Mc/s.

Since it is usual to work in inches rather than feet, particularly so far as the short u.h.f. aerials are concerned, the following formula is useful:

$$l=\frac{5616}{f}$$

where l is the overall length of the dipole, in inches, and f is again the frequency in Mc/s.

Wideband Working

Although an aerial is designed towards a specific frequency the tuned circuit is heavily damped and the effective Q is low. This gives the aerial wide-band characteristics so that signals using frequencies removed from the design frequency are also embraced. It is because of this that a single aerial

serves for both the vision and the sound signals. If an aerial was a high Q circuit and damping was not possible two aerials would be needed for each television channel—one for sound and one for vision!

Aerials designed for the higher frequency channels have a greater bandwidth than aerials designed for the lower frequency channels. For example, a Band I aerial needs to be cut fairly accurately to the required channel, since its response falls off relatively sharply either side of the tuned channel.

A Band III aerial has a wider passband and such an aerial would probably respond to the adjacent channels either side of the tuned channel, thereby rendering it suitable for use over three adjacent channels.

On the u.h.f. bands, the bandwidth of an aerial is automatically greater and most commercial BBC-2 aerials are designed for working over a spectrum of 88Mc/s. In the London u.h.f. region, Channels 23, 26, 30 and 33 are allocated. Channel 33 was the first to start, carrying BBC-2.

Each u.h.f. channel is 8Mc/s wide with guard channels between. Thus a u.h.f. aerial for the London region should be designed to embrance eleven 8Mc/s channels from about 486Mc/s to 564Mc/s. The same will apply to aerials for certain other regions (Suffolk, for instance). However, this 88Mc/s spectrum will not apply in all parts of the country.

Eventually four u.h.f. channels will be available in each local area in accordance with a pattern of channels established by various authorities with regard to the possible use of booster stations, co-channel interference and so forth. In many areas, the four u.h.f. channels available will embrace a spectrum of 88Mc/s—as mentioned earlier, the London area channels cover 486Mc/s to 574Mc/s.

At present (1967), only one channel is active in the areas so far covered by the u.h.f. transmissions, but sooner or later, the four-channel spectrum will be in use. Thus, u.h.f. aerials, pre-amplifiers and tuners, etc., have to be designed to cover all four channels of the area in which they are to be used.

To simplify matters, designers of aerials and amplifiers divide the u.h.f. channels into three groups "A", "B" and "C", which cover Channels 21 to 34, 39 to 51 and 52 to 68 respectively. A system of colour-coding is also adopted: Red for "A"; Yellow for "B"; and Green for "C".

The point to note from all this is that home-constructed aerials should in general be of the wide-band type so that their responses cover all of the four channels eventually to be brought into use in the area.

However, for the very best results, especially in fringe areas, an aerial should be tuned for one particular channel. This also applies to DX television activities.

Nevertheless for most general u.h.f. reception an aerial designed for broadband working over 88Mc/s is perfectly adequate.

The mean frequency of the London-region u.h.f. channels is 530Mc/s, so a London dipole should be cut towards that frequency. The overall length of the dipole is then 5616/530in (from the formula given earlier), which is 10·6in. This is a little longer than a dipole designed for Channel 33.

A single dipole of only about $10\frac{1}{2}$in in length does not represent very much metal with which to "grip" the signal and only in very strong signal locations, free from multipath interference (ghosting), would such a simple aerial work.

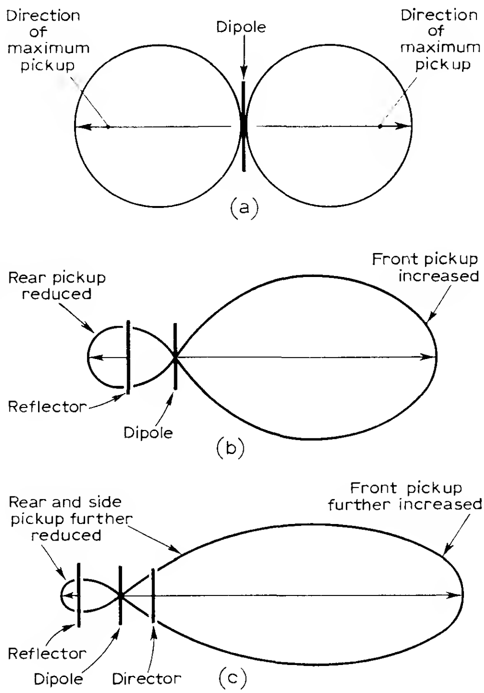


Fig. 9. A single horizontally mounted u.h.f. dipole has a polar diagram as shown at (a), where the pick-up is equal at either broadside position, and minimum along the length of the dipole. The pick-up pattern at (b) is obtained by the addition of a reflector, and that at (c) by the addition of a reflector and director.

Reflectors and Directors

The dipole cannot be increased in length because the tuning would alter, so the only thing that can be done to make the aerial more "powerful" is to add a reflector and a number of directors.

A rod placed behind the dipole is called a reflector and a rod placed in front of the dipole is called a director. A reflector considerably reduces the pick-up at the rear of the aerial and adds to the pick-up efficiency at the front, while a director further adds to the front pick-up efficiency and tends to cut down the response at the side. Reflector and director rods are known as parasitic elements.

Parasitic Element Lengths

Most u.h.f. signals are horizontally polarised. This means that the aerial must be positioned so that its rods are horizontal. A single horizontally

mounted dipole has a signal pick-up pattern (polar diagram) in the form of a figure eight as shown at (a) in Fig. 9. Diagrams (b) and (c) show how the pick-up pattern is altered by the addition of a reflector and reflector plus director respectively.

For the best results, the reflector should be about 5% longer than the dipole and the director about 5% shorter than the dipole. The following formulae are useful in calculating the length of these elements:

$$lr = \frac{5976}{f}$$

$$ld = \frac{5400}{f}$$

where lr is the length of the reflector, in inches; ld is the length of the director, also in inches; and f is the frequency in Mc/s.

For the London-region stations, therefore (with a mean frequency of 530Mc/s), the reflector would be about 11·3in and the director about 10·2in.

Spacing of Elements

For the best gain the reflector should be spaced about 0·15 wavelength from the dipole and the director about 0·1 wavelength from the dipole.

The wavelength of the London-region stations (that is the mean wavelength) is (300/530)m, which works out to 0·565m. 0·15 of this value (0·565 × 0·15) works out to 0·08475m or 8·475cm. To convert centimetres to inches we divide by 2·54, which gives the dipole-to-reflector spacing of 3·3in.

0·1 of 0·565 (0·565 × 0·1) works out to 0·0565m or 5·65cm, which converts to 2·22in and gives the dipole-to-director spacing.

It is in the element spacings that we may find the greatest difference between aerials of different design, even though they may all be designed for the London-region channels.

Reflector-to-dipole and director-to-dipole spacings of 0·15 wavelength and 0·1 wavelength respectively, whilst giving the greatest gain, do tend to be somewhat critical, and vibration of the elements due to wind or other causes could cause picture flutter effects. For this reason, arbitrary spacings, somewhat in excess of those for maximum gain, are often employed in practice. Typical dipole-to-reflector and dipole-to-director spacings are 5in and 3⅜in respectively.

Now when parasitic elements are added to a dipole aerial the impedance at the centre of the dipole (see Fig. 4) tends to fall well below the dipole-only impedance of 72Ω. Indeed the impedance may drop to around 18Ω and a very poor match then exists between the dipole and the feeder cable.

The Folded Dipole

Something must be done to step up the dipole impedance again so that it approximates to that of the cable. One simple and effective way of accomplishing this is by folding a full-wave rod dipole as shown in Fig. 10; this

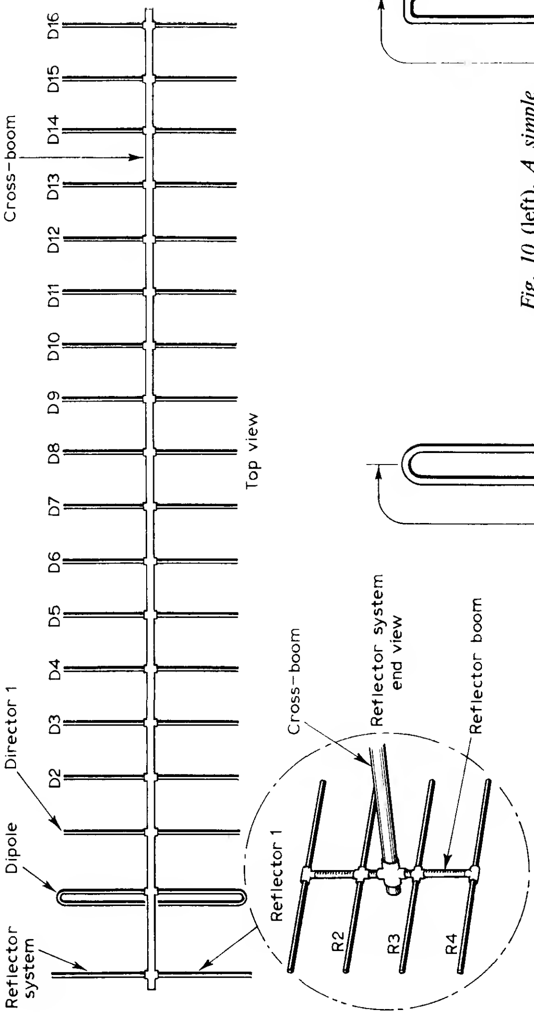


Fig. 12 (above). A 21-element u.h.f. aerial. All the directors are the same length and all the reflectors are equal in length too. Duralumin tubing is suitable for the cross-boom and Duralumin rod (3/16in diameter) is suitable for the elements. The array is best supported at the point of balance along the cross-boom, and it must be orientated for maximum pick-up of signal.

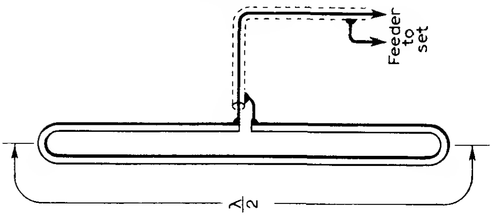


Fig. 10 (left). A simple folded dipole. This increases the impedance four times.

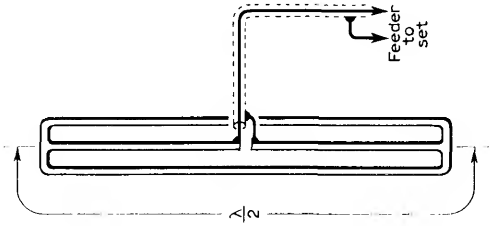


Fig. 11 (right). A triple-fold dipole.

results in a folded (half-wave) dipole. One simple fold like this increases the centre impedance by four times, so if the impedance is down to 18Ω due to the addition of parasitic elements, a simple fold steps it up again to 72Ω. Very convenient, to say the least!

This technique is adopted on the majority of u.h.f. aerials and in some very complicated designs there may be a triple fold (Fig. 11) and tapering of the dipole metal to secure optimum conditions over a wide frequency spectrum. A triple fold incidentally increases the impedance nine times. The length of a folded dipole is measured from the middle of the bend at one end, round one of the sides to the middle of the opposite bend, as shown in Fig. 10.

Typical Arrays

We now have the basic ingredients necessary to make up u.h.f. arrays. As the elements and element spacings are so small, arrays of up to 21 elements are relatively compact. There is not much point in adding any more than about 16 directors (the director-to-director spacing should equal the director-to-dipole spacing originally computed), but with four reflectors, plus the dipole, this gives an array of 21 elements as shown in Fig. 12.

Instead of reflector rods, the reflector can comprise a sheet of expanded metal (alloy for light weight) about a full wavelength square (22¼in square for the London-region channels). Spacing should then be approximately a quarter wavelength from the dipole (5½in for the London channels).

Outside Aerials

Mechanically, an outside aerial needs to be more rugged than its attic-mounted counterpart, though technically there is not a lot of difference between them. At this early juncture, however, the point should be made that at frequencies corresponding to Bands IV and V, the signals passing through the roof of a house are weakened or attenuated several times more than signals of the v.h.f. channels, and this applies more when the roof is damp or wet.

The u.h.f. attenuation is least when the roof is perfectly dry and when the signal arrives at right angles to the roof surface. The worst possible condition for an attic aerial is the centre of a terrace block with the signal passing along the length of the building! It is just not worth considering an attic aerial in such a case.

Neither should attic aerials be considered if there happens to be a lot of foil or metal in the roof space. Sometimes metal foil is employed for heat insulation, and this can prevent the use of attic-mounted u.h.f. aerials.

Generally speaking, the best possible outside aerial should be used in all areas other than those in a very high signal field. Note that even close to the transmitter, outside aerials may be needed to get rid of ghosts or to overcome some local screening problem.

Set-top aerials and other types of aerial installed in the same room as the receiver are not very satisfactory at ultra-high frequencies, and recent reports have been received of signal variations due to the movement of knives and forks when viewing-room aerials have been used. This, of course, is not surprising when it is considered that an average fork is approaching a half-wavelength of the London u.h.f. channel (Channel 33).

Greater liberties can be taken with the v.h.f. serials, and in cases where a roof or chimney stack is cluttered with such arrays, an attempt should be made to bring them indoors (into the attic), thereby making room for the more critical u.h.f. arrays. If the v.h.f. signals fall too much by this technique, a pre-amp can be employed to restore the strength.

Building an Attic Aerial

For attic use, the mechanics of aerials are greatly simplified and a hardwood cross-boom is satisfactory. The small dimensions of the elements and the fact that the u.h.f. transmissions are often horizontally polarised are other features which make attic-mounted aerials a good proposition (always provided the conditions permit their use as emphasised above).

The general disposition of the aerial elements in a Yagi array for Channel 33, and the essential dimensions are given in Fig. 13.

Fig. 14 shows how a length of hardwood batten can be arranged to act as the cross-boom to secure the elements. Small staples, or metal "U"-clamps held in position by wood-screws, can be employed to hold the elements at the calculated points along the length of the batten. A length of 1in × ¼in batten is suitable, but the wood should be thoroughly dry and of good quality.

An alternative arrangement is shown in Fig. 15. Here, the boom is made of a piece of 1in × 1in wood and holes are drilled to accommodate the elements. The diameter of the holes must be such that the elements are a tight fit when pushed through the boom.

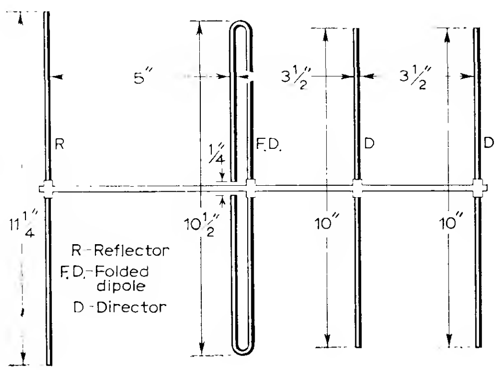


Fig. 13. Lengths and spacings of aerial elements for Channel 33. All elements are made from ⅛in copper tube.

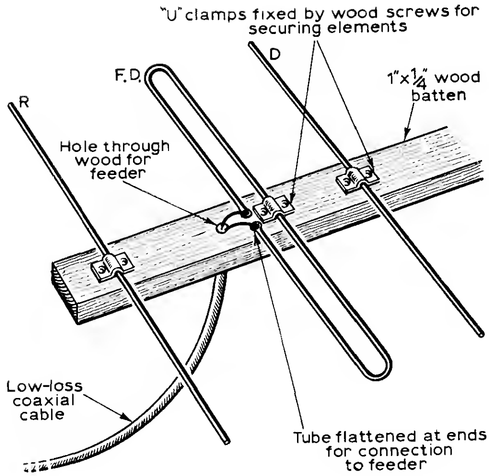


Fig. 14. Practical diagram showing how the elements of an indoor u.h.f. aerial may be secured to a 1in × ¼in wood batten.

For attic use, the elements can be made of ⅛in diameter copper tube. This is readily available from garages and is easy to work with, even if machining facilities are not available.

The biggest problem is in stretching the tube to straighten it so that it can be cut accurately to length. When it is purchased, it is rather kinky and coiled since it is usually rolled. However, clamping one end in a vice and pulling one's weight against it solves the problem, but care should be taken to ensure that the vice itself is adequately secured, and also the tube between the clamps.

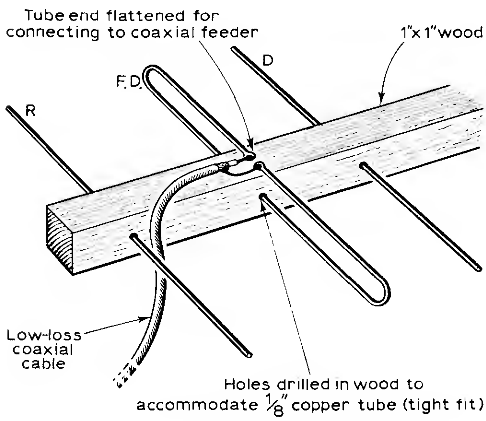


Fig. 15. An alternative method of fixing the elements by drilling through a 1in × 1in cross-boom.

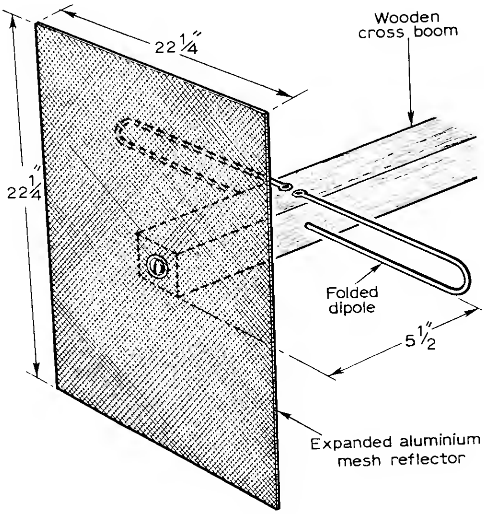


Fig. 16. Expanded aluminium can make a useful reflector and is easily fixed on to the end of a cross-boom.

Making the Folded Dipole

The dipole fold is achieved by shaping the tubing round the handle of a broom, but special attention should be paid to the dimension of the dipole, bearing in mind that the length is measured from the centre of the bend at one end along one side, to the centre of the bend at the other end (see Fig. 10).

It is best to waste a short length of tube for the dipole by making the open ends overlap. The correct distance of about ¼in between the open ends is then easy to obtain, after which the ends should be hammered flat for making the connections to the cables, and for screwing to the cross-boom.

Soldered connections to the inner and outer coaxial conductors are desirable, but before this is done the elements should be thoroughly polished with a mild abrasive. The polishing should, in fact, take place before the elements are mounted on the boom.

After mounting, a final polish should be given and then the elements may be given a couple of coats of the special type of transparent varnish of the kind that is sold to prevent discoloration of polished brass and copper. Note that this substance is highly inflammable and should be applied in a well ventilated room clear of open flames. The first coat must be dry before the second coat is applied. The flat ends of the dipole should not be coated, of course, until the coaxial cable has been connected.

It was mentioned earlier that instead of a rod or a series of rods, the reflector can consist of a sheet of metal mesh. The reflector should be at least one wavelength by one wavelength. A suitable material is expanded aluminium

of the kind that is often employed for loudspeaker grilles. This is easily fixed to the end of the wooden boom, as shown in Fig. 16. The dimensions given in this diagram relate to Channel 33.

"Bow-tie" Aerial

Another attic type aerial suitable for the u.h.f. channels is the so-called "bow-tie" or "batswing" dipole used extensively in the U.S.A. and in Europe. Fig. 17 gives all the measurements and constructional details for building this aerial for use on Channel 33. The information given earlier will enable the design to be altered to suit other channels.

The reflector is designed for a 60° angle and this is initially established by the critical dimension of the wooden dipole-support. Stability of the reflector at the correct angle is then achieved by the wooden supports at the rear. The distance of the "bow-tie" dipole from the corner of the reflector is rather critical, as this establishes the gain and the dipole impedance. It will be seen that the dimension also has a bearing on the reflector angle.

The reflector proper can be made of expanded aluminium or of sheet aluminium, the latter being more rigid and the former probably calling for some form of outside wooden framework.

Any type of dipole can be experimented with, but that shown is rather interesting. It is made of flat aluminium of an overall length of 10½in, which allows for a ¼in spacing between the two half sections. The narrow ends are ½in wide and the outside ends 2in wide. The necessary bandwidth is achieved by virtue of the unusual shape of the dipole.

(The length of the bow-tie dipole is basically that of a conventional rod type, but as the width or diameter is increased, the length should be correspondingly decreased. It will be remembered that increasing the width or diameter of a dipole increases the bandwidth.)

An ordinary 10½in (overall) rod dipole can be used almost equally as well, but a fold is not necessary for impedance matching provided the spacing of the dipole from the corner of the reflector is as shown in Fig. 17.

This type of aerial has a very high gain and front-to-back ratio, and is ideal for attic mounting, provided it is not "looking" along the length of a terrace block of houses.

Outdoor Aerials

Aerials for outdoor erection follow the same general principles as described for attic types, but greater stability to withstand the weather is essential. Moreover, high gain u.h.f. arrays are extremely directional, which means that picture flutter could be bothersome if the aerial vibrates or moves with the wind.

For outdoor arrays, the special components available to aerial constructors should be employed, for these days it is hardly worthwhile

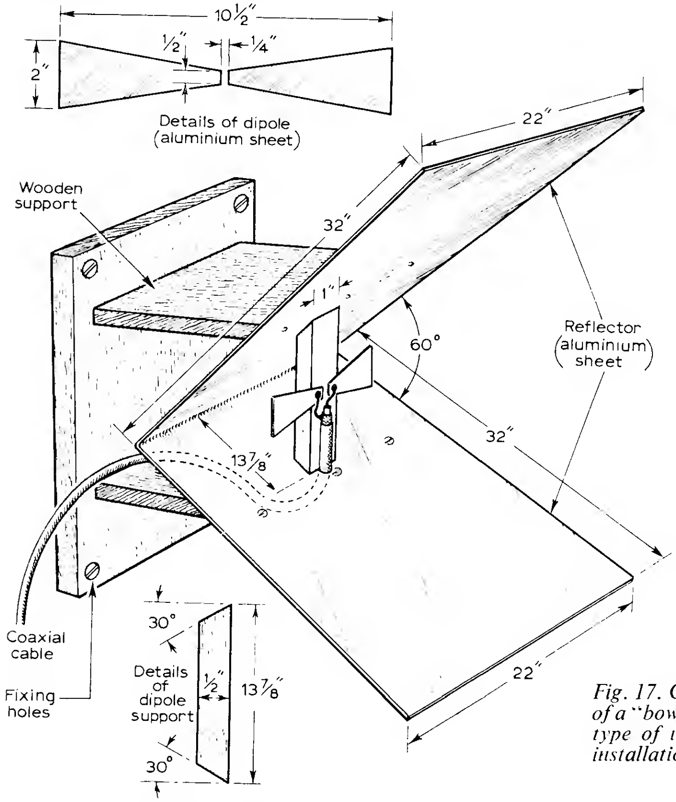


Fig. 17. Constructional details of a "bow-tie" corner-reflector type of u.h.f. aerial for attic installation.

machining special rods, insulators, brackets and so forth as they are so easily obtainable to almost any requirement.

There is little doubt that the greatest scope for the experimenter lies in the design of attic-mounted arrays, such as have been dealt with in detail here.

As a final thought, it is as well to check the house insurance before mounting home-made aerials on chimney stacks.

Cables for U.H.F.

Normal coaxial cable is likely to prove unsatisfactory for Bands IV and V. At higher frequencies, transmission losses in cables are increased and cause serious falling-off in signal strength. The importance of using suitable coaxial cable for aerial downleads especially in "fringe" areas cannot be overstressed.

Special cables have been designed by several manufacturers for Band IV and Band V applications. These cables have larger inner conductors and improved polythene insulation to reduce the rate of attenuation at ultra high frequencies.

Chapter Eight

A DX AERIAL SYSTEM

NOT only can a great deal of interest and excitement be derived from the reception of distant television transmissions but with the advent of the dual-standard television receiver a combination of European transmission standards can easily be catered for.

Great Britain is within the "tropospheric propagation" distance of a large number of television transmissions originating on the other side of the English Channel. During the early summer months, peaking again during late autumn, v.h.f. television signals are carried well beyond their "local" range by a "ducting" effect in the upper atmosphere or the "troposphere".

Due to this, stations which are normally receivable over a maximum of 60 miles or so come in over distances of several hundreds of miles. This is good for the DX (long-distance reception) fan but unkind to the non-enthusiastic viewer, since it can badly upset his local reception, giving drifting ghost pictures, horizontal dark lines and a Venetian blind effect on the picture, and a loud buzz on the sound.

Tropospheric propagation occurs mostly on the Band I channels, but is sometimes present on the lower Band III channels. Reception of stations up to 500 miles distant is possible by this means, but even greater distance reception is possible by so-called sporadic E skip. Here, instead of the troposphere, the E layer of the ionosphere is concerned with the freak reception. This propagation is primarily tied to the lower frequency channels of Band I, and can give receptions over distances from 500–1500 miles (sometimes greater distances by a "double hop" effect).

When the DX enthusiast observes the first signs of freak reception on the local stations his mind turns to ways and means of isolating the DX signal from the signal of the local station and then arranging his receiver to respond to it.

There are two basic requirements: (1) a good, high-gain aerial system tuned to the mean of the distant channel and orientated for maximum signal pick-up and (ii) a highly sensitive receiver with a very low noise figure. The receiver, of course, must be suitable for the DX transmissions, and information on the characteristics of European television stations is given in a number of books.

Theoretically, DX reception (and hence "co-channel interference") is not possible on the u.h.f. channels as such high-frequency signals penetrate both the troposphere and ionosphere and are lost in space without spoiling local

reception, irrespective of the weather and sun-spot conditions. This is one of the reasons why America and some of the European countries are contemplating the abandonment of television in the v.h.f. channels in favour of a total switch to the u.h.f. channels. However, this does not always work out in practice, and with good tropospheric conditions DX on u.h.f. is certainly a factor to be borne in mind.

The Aerial

Under severe conditions, where a strong local station is interfering with a DX signal, traps and filters may not succeed in cutting out enough of the local signal to allow the distant station to be viewed or photographed. This is because adequate cancellation of the unwanted signal would demand many more decibels of attenuation than simple filters could possibly supply.

However, a method of cancellation of the unwanted signal is possible by the use of two aerials. The main, high-gain aerial is carefully tuned and orientated to pick up as much of the DX signal as possible, while the second aerial is arranged to pick up as much of the interfering local signals as possible and the smallest amount of the DX signal.

The idea, then, is carefully to adjust both the phase and the amplitude of the picked-up local signal so that it is of equal amplitude and of opposite phase to the unwanted signal picked up by the main aerial. In that way, complete cancellation of the local unwanted signal is theoretically possible. In practice, a very useful degree of unwanted signal attenuation is possible (see Fig. 1).

The two aerials are connected to a common downlead through a star network, but the second aerial signal is first passed through an attenuator and "phaser". As the main aerial will be orientated for maximum pick-up of the DX station, a certain degree, even though it may be small, of discrimination over the strong local signal will be achieved.

This means that the second aerial need not usually be so complicated as the main aerial, for the requirement here is that this picks up more local signal than the main aerial, and since it can be pointed direct to the local station this is rarely difficult to achieve, even with an ordinary "H" array or, in some cases, with a single dipole.

Amplitude Adjustment

It must be stressed, however, that the system will not work unless the second aerial picks up a little more unwanted signal than the direct aerial. The two unwanted signals are adjusted for equal amplitude by the attenuator in the downlead from the second aerial.

Reasonable signal balance is possible by the use of Belling & Lee plug-in attenuators, but for optimum results a variable attenuator can be used. The range of attenuation will, of course, depend upon how much stronger the local signal is from the second aerial than the same signal from the main aerial.

If the second aerial signal is twice as strong, for example, a 6dB attenuator will be required; if three times as strong 9dB attenuation is required, and if four times as strong 12dB is required. The better the signal balance, the better will be the cancellation of the unwanted signal.

The biggest problem in setting up this system is, in fact, adjusting for signal balance. A signal strength meter solves the problem, for then the signal in the main aerial downlead can first be measured, and the signal in the second aerial downlead adjusted by the attenuator for an equal value. The "TV S-Meter", described in Chapter 25, page 313, can be employed for this exercise.

At a push, the two signals could be compared in terms of picture and sound on a receiver, but care should be taken if this method is adopted as the a.g.c. action of the set may tend to make both signals appear to be of similar levels, even though they may in fact differ by several decibels.

Phasing Adjustment

Once the signal levels are balanced and the arrangement set up, minus the phaser, as shown in Fig. 1, it is quite likely that a good amount of unwanted signal discrimination will have automatically taken place.

The reason for this is that the phase of the unwanted signal in the second aerial may be almost opposite to that of the unwanted signal in the main aerial. The phase of the signals will depend upon the standing wave conditions around the aerials, how well the aerials are matched to their downleads and their positions relative to each other. It is unlikely that the signals will be either "in phase" or "anti-phase".

But if they are in phase the unwanted signal will be much stronger than on the main aerial alone, while if they are in anti-phase the unwanted signal will be at a very low level.

It is likely that the phasing will be at some intermediate value, indicated by the presence of still a fair amount of unwanted signal plus some deterioration in quality of the wanted picture. One simple way of securing the correct anti-phase condition is by a trial and error method.

This involves cutting the cable between the attenuator and the star network inch by inch until the right length for unwanted signal cancellation is established. If an extra 4½ft of cable is left at that point, the trial and error method should not give a great deal of difficulty.

Some simple cable clamping arrangement can be fitted to point "C" of the "star" so that after cutting, the cable can quickly be connected again, and so on. When the best conditions have been established a more permanent connection can be made.

If the second aerial picks up some of the wanted signal, the quality of the DX picture may suffer a little after setting up the scheme for maximum unwanted signal rejection. This is due to phase distortion introduced by having two *wanted* signals in the system. If necessary, the second aerial

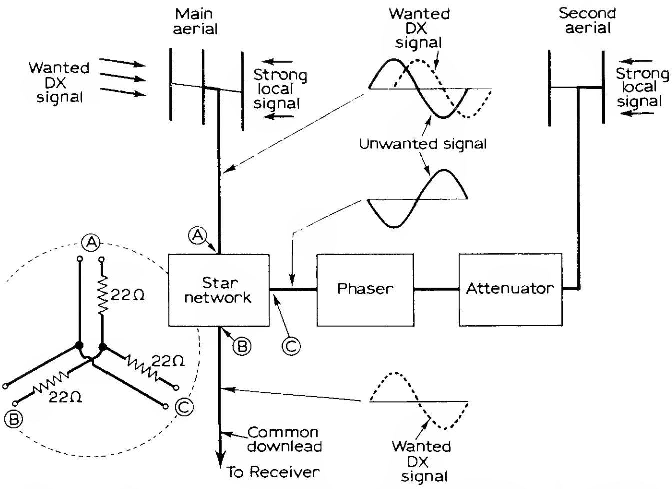


Fig. 1. By the use of two aerials, a local signal can be cancelled leaving only the wanted DX signal. The main aerial is carefully turned for maximum pick-up of the DX signal while the second aerial is turned for maximum pick-up of the local signal and minimum pick-up of the DX station. After balancing of levels and phase-correction (as described in the text) the two signals are combined in a common downlead where the unwanted local signal is cancelled.

should be orientated for the *minimum* pick-up of the *wanted* signal, as distinct from the maximum pick-up of unwanted local signal.

It will be appreciated that the system described has a very great experimental potential, and apart from its applications for DX television it could also be experimented with for getting rid of co-channel interference, when the emphasis is on local reception rather than distant reception.

Chapter Nine

A PRINTED CIRCUIT AERIAL FILTER

SEVERAL kinds of interference commonly affect the reception of 405-line television pictures. They include the following, apart from ignition noise and mains-borne interference from brush-fed motors:

(a) Signals received at, or within, 3Mc/s of the intermediate frequency of the receiver.

(b) Signals received near the signal-frequency.

(c) "Image" signals from external sources.

(d) I.F. harmonics, produced at the detector, being fed back to earlier stages.

(e) Oscillator harmonics generated in the receiver itself.

Elimination

Strictly speaking, there should be no difficulty in eliminating all these sources of interference. Good filtering and screening of the receiver itself is, in any case, necessary to remove the last two in the list—and perhaps decoupling more than screening. Reception via the aerial of interfering signals is dependent on several factors. One of the more important is that no attempt is usually made to overcome the lack of "balance" that exists between aerial and coaxial feeder. As a result, the coaxial lead itself acts as an aerial and although it is very inefficient at picking up television frequencies, whether Band I or Band III, it is liable to receive a good deal of short-wave energy. Moreover, the lead comes down from the aerial through a region where—especially in towns—other television receivers are working and are generating as a rule a good deal of radiated energy.

While the use of suitably-proportioned balance-to-unbalance transformers ("baluns") could be used, it is normally much simpler to dispense with this complication and use other means of overcoming the trouble.

Trap Circuits

If a single local source of radiation is the cause of "patterning" on the received picture, a common device is to include somewhere in the lead—and

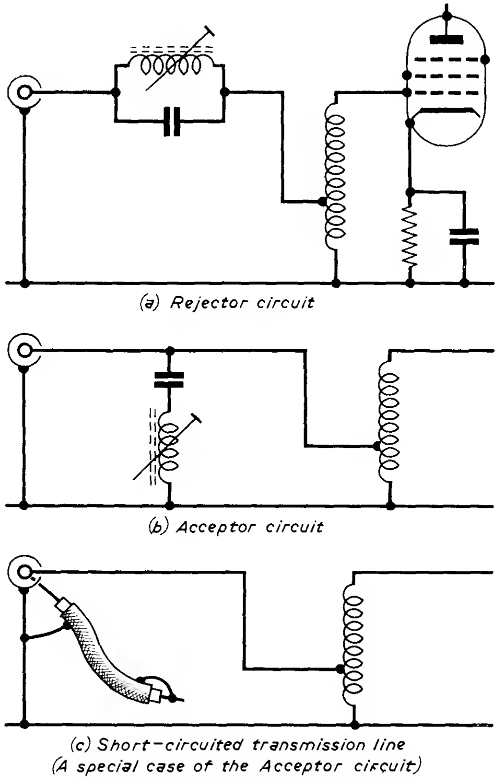


Fig. 1. (a) Rejector circuit; (b) Acceptor circuit; (c) Short-circuited transmission line (a special case of the acceptor circuit).

as near as possible to the receiver—a simple trap circuit. Three possible arrangements are shown in Fig. 1.

Usually about 30dB attenuation of an interfering frequency is aimed at, and although the arrangements shown in Figs. 1a and b are sometimes used, only Fig. 1c will give this degree of attenuation in normal circumstances.

Where several different interfering frequencies may exist, none of these circuits is satisfactory. This is usually the case—the considerable bandwidth of the vision i.f. amplifier allows for many different frequencies to enter the amplifier. Under these conditions it is very difficult to remove unwanted frequencies by simple trap circuits, largely because the addition of extra traps affects those already fitted. The alignment is very tricky and the difficulty increases geometrically as the number of traps increases.

Printed Filter

The best method is the use of a suitable wave filter, and here is described the construction of such a filter using printed circuit techniques. The arrival

on the "home experimenter" market of "do-it-yourself" printed circuit kits makes this feasible—although the amateur needs only to be able to obtain the copper-clad insulating board as the chemicals needed are freely obtainable and very cheap.

The wave filter is a combination of inductances and capacitances so chosen that, when inserted in a circuit between a generator and a load, very great attenuation of certain frequency bands occurs with negligible effects on other ranges of frequency. If the calculations have been successful, a very abrupt transition occurs between the wanted and unwanted frequencies.

Matching

There are three types of filter in general use: the low-pass, the high-pass and the band-pass. Here we are concerned with a high-pass filter—one which eliminates all frequencies below 41Mc/s and transmits, with little loss, all frequencies above this.

Any filter must be designed for definite input and output impedances. The filter described here is only suitable for inclusion in a coaxial line of 75Ω nominal impedance.

Wave filters normally incorporate both end-sections and intermediate-sections. The former are concerned chiefly with the achievement of the correct input and output impedances; the intermediate-sections supplement the attenuation given by the end-sections, allow for very great attenuation at certain frequencies, and ensure more uniform attenuation over the whole range.

"π" or "T" Types

Filters may be either π-type or T-type. This means, roughly, either of rejector or acceptor type. For this design, the π type has been chosen, because the self-capacitances of the coils affect the behaviour less, in that the capacitors added in the design are in parallel with coil self capacitances. The theoretical circuit is shown in Fig. 2. Although one end is labelled

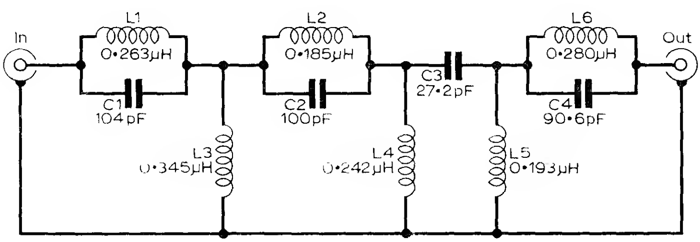


Fig. 2. The theoretical circuit of the filter.

"In" and the other "Out" they are in fact interchangeable with no effect on performance.

Two possible methods of construction offer themselves as attractive alternatives. The printed circuit system has the advantage of simplicity of execution, a minimum of capacitance between input and output ends and reduction of coupling between the inductances. The alternative, of using coils wound on small formers, enables more accurate adjustment, especially of the very sharp edge of the transmission band. However, this makes alignment rather more difficult. For those who prefer not to use printed circuits, details of the coils will be given and of an alignment method. It should be noted that in this case screening of the whole unit will probably be required.

Construction

The copper-clad boards supplied with "do-it-yourself" kits usually measure about $3\frac{1}{2}$in × 6in and this is a convenient size to use. The first thing to do is to clean the board with kitchen scouring powder, so that the copper is clean and bright. Then a line is drawn with a soft pencil 1in from each long edge of the board, right across it. Marking out and drilling is then completed as in Fig. 3.

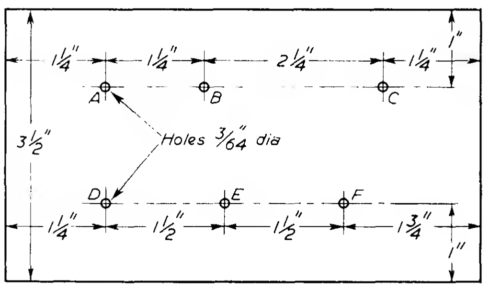
A brass pin of normal size (1in long) is required, and it should be examined under a hand lens to ensure that its diameter is just a shade under $\frac{1}{32}$in (i.e. 22 s.w.g.).

A length of No. 40 Machine Twist is attached to this pin by a knot and is secured to the metal firmly with a touch of waterproof cement. When this is thoroughly dry the cotton is cut off to a length of about 2in and a small loop about $\frac{1}{8}$in diameter is tied at the end, to take a pencil point.

Marking the Coils

The copper-clad board, drilled with holes as shown in Fig. 3, is placed on a wooden surface. The pin with cotton attached is pushed through a hole into the wood so that it is firmly held. A soft (4B) pencil point is put through the prepared loop in the cotton and marking off the coil can begin. The idea

Fig. 3. Marking out the board.



is to keep the cotton uniformly tight while winding it on to the pin, meanwhile marking out the coil. The winding-up (see Fig. 4) is done until the pencil point comes to within about $\frac{1}{8}$in of the pin and the end of the coil is fashioned as shown in Fig. 5, so as to obtain a definite point where the coil begins. Counting from the centre the turns for each coil are as follows:

Coil Centre	*No. of Turns*	*Approximate maximum outer diameter*
A	$4\frac{1}{2}$	0·65in
B	$4\frac{1}{4}$	0·6in
C	$4\frac{3}{4}$	0·675in
D	5	0·7in
E	$4\frac{1}{8}$	0·65in
F	4	0·6in

The figures for the outer diameter are given to afford a check on the dimensions of the spiral; if markedly more or less than this, a pin of different diameter or other cotton should be chosen.

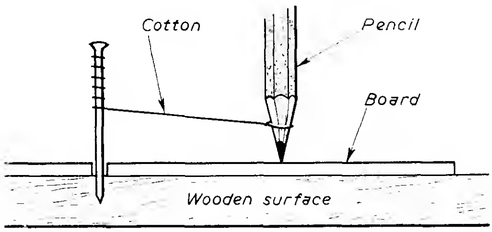


Fig. 4. Marking out the coils.

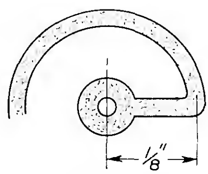


Fig. 5. Details of the centres of the coils.

"Resist" Paint

When the spiral coils are marked in pencil they have to be covered with the "resist ink", which consists of black cellulose paint. With care, and the use of a fine brush, a steady enough line can be painted over the top of the pencil line already marked out. The thickness of the paint line is so arranged as to give, roughly, a spacing between turns equal to the width of the line, or a little over.

When dry, the paint line can be tidied up with the point of a penknife blade if any major inaccuracies have occurred in drawing. More of the copper is now covered with the cellulose paint so that not only is there sufficient copper left, after etching, to give low r.f. resistance to the "chassis" but also some screening is afforded between coils. Reference to the illustration (Fig. 6a) will indicate the sort of result to be achieved. It is advisable at this stage also to allow for holes to be etched to take coaxial sockets for input and output. The "chassis" must not be brought too close to the coils.

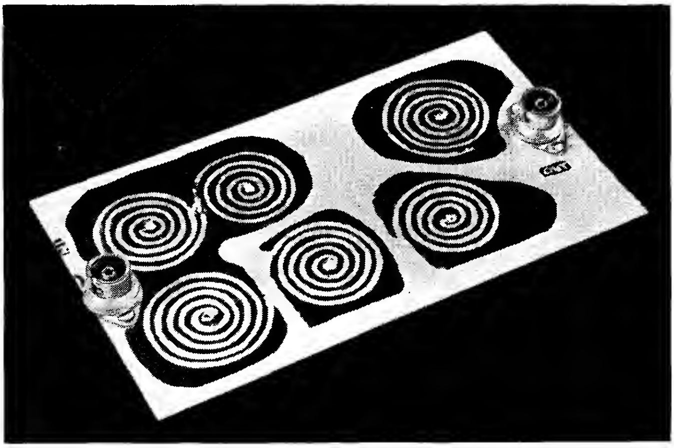


Fig. 6a. The complete "printed" circuit.

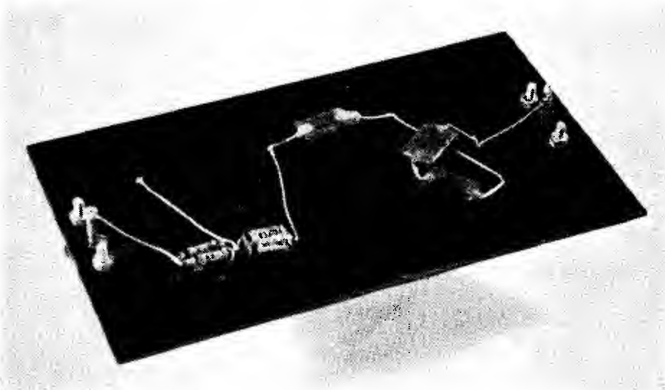


Fig. 6b. The reverse side of the printed board. The series/parallel combinations of the capacitors used to achieve the necessary values of capacitance can be clearly seen.

Cleaning

When the board is completely dry and the paint hard it should be washed thoroughly in hot detergent solution. Grease from the fingers may otherwise cause uneven etching. It is then placed in the etching solution, which is a strong solution of ferric chloride in water, and after about half an hour the copper not covered by the "resist" will have been etched away.

The cellulose paint can now be scrubbed off with kitchen scouring powder. Cellulose "thinners" are unnecessary for a small job like this, and are very dangerous if inhalation of the vapour occurs. For best results, at this stage the remaining copper should be silver-plated.

Other Components

The capacitors used may be ceramic or silver-mica. The important thing is that they should be of 5% tolerance or better. Best results will be achieved if facilities exist for accurate capacitance measurement, when values can be selected from the constructor's stock to the nearest 1%. The difficulty exists that the edge of the pass-band is very close to the sound frequency of Channel 1, and so London viewers may find that some attenuation of 41·5Mc/s occurs, unless care is taken in selecting capacitances. If this happens, C2 and C3 should be varied a little each way (say, 3–5pF) until the desired attenuation characteristic is achieved. The illustration of the reverse side of the unit (Fig. 6b) shows how in the prototype series or parallel connection of capacitors was used—this was because higher accuracy was thus obtainable. The characteristic obtained with the prototype is shown in Fig. 7, and it will be seen from this that the usual i.f. range of 34–49Mc/s is severely attenuated.

Even this filter gives no protection against "image" (second-channel) interference which may occur by picking up the radiation from f.m. receiver oscillators. Only reasonably good selectivity ahead of the i.f. amplifier can prevent this. An interference signal within a megacycle or two of the desired signal is likewise not affected except that Channel I viewers see and hear no trace of signals below about 40Mc/s.

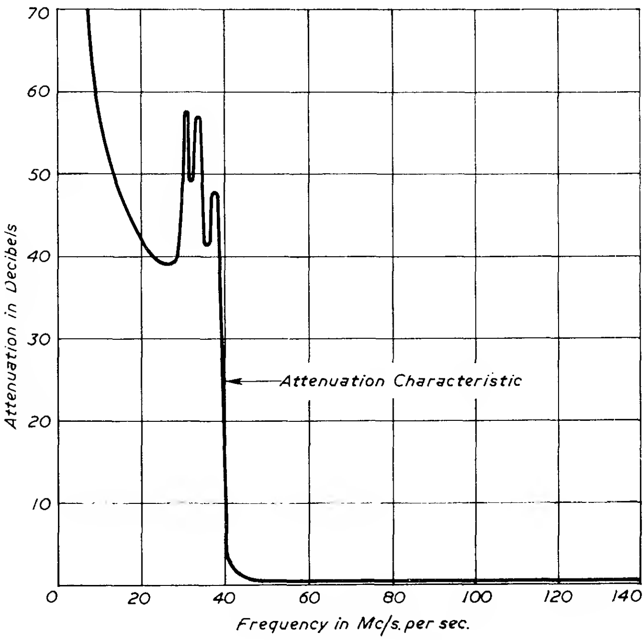


Fig. 7. The attenuation-characteristic of the filter.

The Filter using Conventional Coils

For those who prefer to make up the unit using conventional coils, the following notes will be useful.

All coils except L3 (which has four turns) consist of $3\frac{1}{2}$ turns of 30 s.w.g. enamelled copper wire, close wound on a former $\frac{1}{4}$in in diameter. A v.h.f. type (purple-coded) iron dust slug is provided for each.

To set the coils to the required inductance by adjustment of the core, a 100pF capacitor (1% or 2% accuracy) is placed across the probe terminals of a valve voltmeter. Using the shortest possible connector leads the valve voltmeter is attached across each inductance in turn. A signal generator is coupled very loosely indeed to each coil—the "high output" terminal is best used if one is provided—and a short lead from it put 2in or 3in away from the coil concerned.

The signal generator is set to the appropriate frequency, from the following table, and the core of the inductance in question is adjusted to resonance. This is repeated for each coil in turn, and on completion the unit is ready for use as soon as connections have been soldered up.

It should be noted that the filter must be connected as near as possible to the input socket of the receiver. Its efficiency if used properly is very great, but if a long length of coaxial cable joins the unit to the receiver unwanted signals can be picked up on this length and fed straight into the receiver.

ALIGNMENT FREQUENCIES
(loading capacitance of 100pF)

L1	0·263μH—31·1Mc/s	L2	0·185μH—37·0Mc/s
L3	0·345μH—27·0Mc/s	L4	0·242μH—32·2Mc/s
L5	0·193μH—36·25Mc/s	L6	0·280μH—30·0Mc/s

Chapter Ten

AERIAL ATTENUATORS

It frequently happens that the aerial signal, if fed direct to the input terminals of a television receiver, overloads the r.f., frequency-changer, or the i.f. valves, or produces too much "drive" at the cathode ray tube. With modern receivers, effective automatic gain control is usually provided and the problem is not too severe. However, there are many occasions when too much aerial signal is not a good thing, and this is especially true when—for the purpose of cutting out "ghosts" or reducing interference—a high-gain aerial has to be used well within the service area of a transmitter. It will be realised that if an aerial is picking up interfering signals from a direction not that of the transmitter, an improvement will be obtained by greater directivity, but this may well reduce the interference and increase the signal, even to the point of annoyance!

Interference Problem

The problem then simplifies itself to that of reducing the voltage available at the receiver input and with it obtaining a much reduced interference signal. To do this effectively, however, certain conditions must be fulfilled.

Unwanted Effects

First, serious mis-match must not be introduced between feeder and input terminals, because not only will this reduce the amount of power transferred from the aerial to the input circuits (the object of the exercise) but will cause more than one unwanted effect.

The first of these is an increase in noise level. Nowadays the need is fully recognised in designing a television receiver to arrange for the input matching to be for minimum noise rather than for maximum signal. This is especially the case for Band III; for Bands IV and V, the requirement is equally exacting, if not more so.

The second undesirable effect is that of loss of resolution. A simple qualitative analysis will show how this may occur. If the mis-match between feeder and input exceeds a certain quantity, an appreciable amount of energy will be reflected back from the input terminals and will travel back to the aerial. If the aerial is well matched to the feeder, this energy will be absorbed and nobody will be the worse—except perhaps a neighbour who receives a delayed signal as well as the desired one and whose picture may thus conceivably be

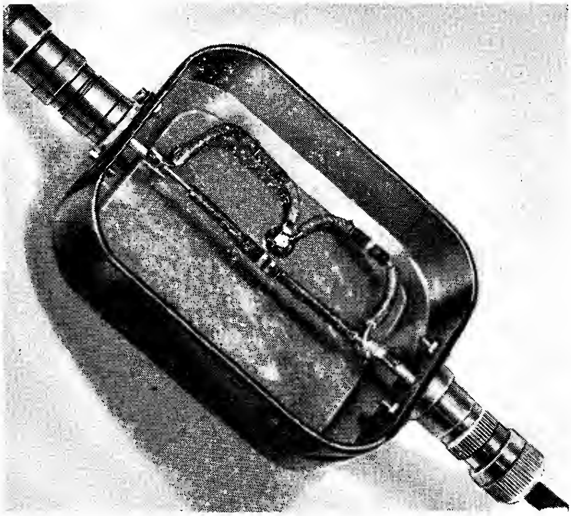


Fig. 1. This clearly shows the simplicity of an attenuator. Inside is the layout of the components for a π-section attenuator.

impaired. Of course this is unlikely to be serious unless perhaps his aerial is only a few feet away from the radiating one.

The more likely case is the effect on the receiver. If the match between feeder and aerial is exact, no reflection takes place at the aerial. This condition is seldom achieved except where a plain dipole is used. Where directors and reflectors are involved there is usually some mis-match in commercial practice. This will involve a further partial reflection, and the re-reflected signal will travel again down the coaxial cable and arrive at the input terminals once more. Further reflection at this point and at the aerial can be neglected because by now the reflected signal will be much weakened.

The total length of coaxial cable so far traversed by the signal may now be 180ft. Although normally by now it will be much weaker than the signal which has not suffered reflection at all, if severe mis-match occurs it may produce a visible signal on the tube face. If it does, it will be delayed about $\frac{1}{3}\mu$s and resolution of the 2·5Mc/s and 3Mc/s bars on the Test Card will be impaired, perhaps lost altogether.

A further effect will be that of changing the damping imposed by the aerial on the input circuits. In correct design, the natural damping caused by the total input resistance of the r.f. valve and its circuits is just about doubled. Where mis-match occurs the damping will be increased or decreased. If the former, noise will increase; if the latter, phase distortion, or reduction of bandwidth, or both, will occur.

T and π Type Attenuators

Any attenuator placed between aerial and input must therefore introduce the minimum amount of mis-match, and this feature has to be kept

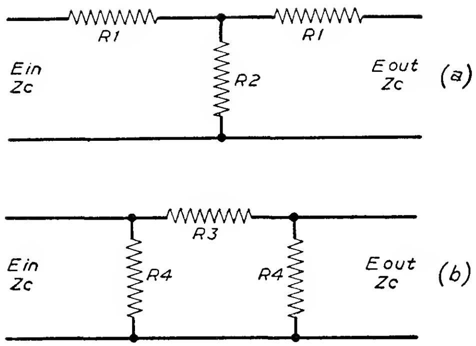


Fig. 2. (a) Circuit for a T section attenuator; (b) Circuit for a π section attenuator.

constantly in mind in design. The most usual feeder is the coaxial, of 80Ω nominal impedance, and this is an "unbalanced" structure. The choice is between T type and π type attenuators, as shown in Fig. 2.

T and π Section Constant Impedance Attenuator

Tables A and B give design data for values from 1dB to 10dB of voltage attenuation. Both T and π sections are included, but it must be remembered that this is for 80Ω nominal impedance only. For other values of feeder characteristic impedance, Zc, the resistor values must be multiplied by Zc/80.

It will be observed that only seldom do the resistors approximate closely to "preferred values". While this would be important from a manufacturing point of view, and would restrict the range of attenuators available, it matters little to the home constructor who can select, from his stock, values within 10% or so of those listed.

It is unwise to attempt to obtain more than about 10dB attenuation in one attenuator. This is because of the inevitable capacitances and inductances associated with practical components, which introduce unwanted coupling between input and output. Instead, where attenuation in excess of 10dB is required several sections can be cascaded.

Constructing an Attenuator

In constructing an attenuator, the output should be separated as far as possible from the input. At the same time, the smallest possible physical size of resistor should be used, because although it brings the output terminals nearer to the input terminals, capacitances and inductances will nevertheless be reduced. If the attenuator is to be a permanent fixture it is also best to avoid the use of coaxial sockets and plugs; these have appreciable capacitance and are often badly matched to the cable. However, where only a temporary component is needed (as in experimental work), they can be used as long as care is exercised.

TABLE A—T section
(Resistors as in Fig. 2a)

	Decibels Attenuation									
	1	2	3	4	5	6	7	8	9	10
R1(Ω)	4·6	9·2	13·7	18	22·4	26·6	30·5	34·5	38	41·5
R2(Ω)	690	345	226	168	131	107	90	75	65	56·2

TABLE B—π section
(Resistors as in Fig. 2b)

	Decibels Attenuation									
	1	2	3	4	5	6	7	8	9	10
R3(Ω)	9·3	18·5	28·2	38	48·7	60	71·5	85	98·5	114
R4(Ω)	1390	700	467	365	285	242	210	185	169	154

An actual case is considered. It is desired to reduce the amplitude of a signal by a factor of $\frac{1}{2}$, or 6dB. The characteristic impedance of the low-loss cable to be used is 75Ω. A short calculation shows that, for a T section (Fig. 2a), R1 is 25Ω and R2, 100Ω. Both these types are readily obtainable—preferred values are 27Ω and 100Ω which are near enough. Had a π section been chosen (Fig. 2b), R3 would be 56Ω and R4, 225Ω—again, close to preferred values of resistor. Here the choice may be made according to what resistors are in stock.

A further practical case, using 80Ω cable, may be taken. Suppose an attenuation of 10dB is required, accurately: the choice is between a T section with resistors of 41·5Ω and 56·2Ω and a π section with resistors 114Ω and 154Ω. Here the former would be chosen, because it would be easier to select from stock (preferred values 39Ω and 56Ω as compared with 120Ω and 150Ω).

Multiple Section Attenuators

Sometimes it is desired to arrange a variable attenuator, operated by a switch. The "ladder" attenuator results from cascading a number of T or π sections, as shown in Fig. 3. However, this simple type is characterised

by considerable variation of input impedance, as will be seen from the diagram. The output impedance also varies, but to a lesser degree.

The resistance values correspond to those in Table B, R5 being given by (R4. Zc/R4) + Zc.

An improvement in the constancy of impedance is obtained by inserting a resistor of value Zc/2 in series with the switch.

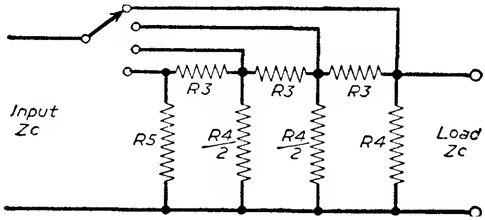


Fig. 3 (left). *π-section ladder attenuator.*

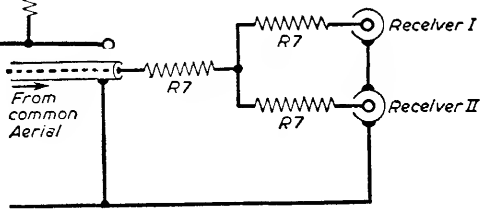


Fig. 4 (right). *A "splitter" circuit for an aerial common to two receivers.*

An Attenuator for Two or More Receivers

Another use of a resistor network exists where several receivers have to be connected to the same aerial. A moment's consideration will indicate that if two receiver inputs are paralleled, the impedance presented to the feeder cable is half that of either. If three receivers, the effective load on a feeder of characteristic impedance Zc is Zc/3. This will cause serious mis-match and can prevent receivers operating correctly. All that is required to overcome the difficulty is to equalise impedance, by means of series resistors, as shown in Fig. 4.

The calculation of R7 is as follows. Assuming all characteristic and input impedances are Zc, the impedance at either receiver input, with the aerial feeder disconnected, is R7 + R7 + Zc.

The aerial feeder will see an impedance, for n receivers, of $R7 + [(R7 + Zc)/n]$ and for correct matching this is equal to Zc.

$$\text{Thus } n.\text{R7} + \text{R7} + \text{Zc} = n.\text{Zc}$$

$$\therefore \quad \text{R7} = \frac{\text{Zc}\,(n-1)}{n+1}$$

Thus, for two receivers with Zc equal to 80Ω, the resistors R7 must be 27Ω which is a preferred value in the ±10% range of resistors.

The signal present at each receiver is, however, reduced by half, and so this type of splitting is only really possible when plenty of signal is available. If several receivers have to be fed from a common aerial in a fringe area it will be preferable to use a pre-amplifier or a much better aerial, or both.

Section 4

MISCELLANEOUS

Chapter Eleven

MAKING PRINTED CIRCUITS

Most experimenters have at one time or another admired the excellent appearance of commercial printed circuits, and most understand the commercial reasons for their introduction. It is probable that some of these reasons have commended themselves to the amateur, notably the advantages of high reproducibility and reliability as well as the ease of working which can hardly be regarded as a vital manufacturing concern. Some may have tried their hand at using the technique for the construction of home-made apparatus, and doubtless many—like the writer—have been rather dismayed by the amateurish appearance of the finished product. However, the advantages of printed circuits are such that it was felt necessary to evolve some simple technique which would put the production of reasonably neat circuit boards within the scope of the experimenter. This chapter offers the results of experiments in this direction. It is not claimed that the last word has been said, however, and any reader may well be able to make improvements.

Principles

The most widely-used circuit board consists of a plastic sheet to which is bonded a layer of copper foil. The insulating plastic sheet is usually a phenolic-paper board the thickness of which varies with the intended use; the most useful gauges for normal work lie between $\frac{1}{32}$in and $\frac{1}{8}$in, and about $\frac{1}{16}$in is probably the most popular size. In addition to phenolic-paper laminate, various other types of board are sometimes recommended for use: epoxide-glass-fibre has excellent electrical qualities, a low absorption of moisture, and can withstand relatively high temperatures; Melamine-glass is useful where a wearing surface is needed, as in switches; and Teflon-glass can be used in microwave circuits where losses are very much to be avoided. Naturally the cost of these special laminates is much higher than the phenolic-paper type.

Printed circuit construction consists essentially of the removal of as much copper foil as is needed, so as to produce an insulated board covered with a network of conducting copper lines or surfaces. Since all is in one plane—or preferably nearly so—some limitations in design have to be accepted. However, with care and the exercise of some imagination the limitations imposed need only be small.

Design

It could be argued that the design of a printed circuit apparatus is chiefly dependent on the artistic or other conception of the finished article. Certainly such practical matters as the placement of pot shafts, tuning condenser spindles, transformers and other heavy items have to be considered from the start if good appearance and accessibility are thought necessary. As a result, it is usually necessary to decide at the outset what the important factors are in the particular task in hand. When these have been well thought out, a certain embryo pattern will emerge which may then be used as a basis for further thought. At this stage the electrical factors must be investigated, and it is as well to list these.

The frequencies of the currents to be carried will govern the type of circuit board selected, the length of the conductors concerned and their spacing. The need for shielding will also have to be taken into account, especially where r.f. or i.f. transformers are to be used. The distributed capacitance of the proposed circuit elements will also be a factor.

The voltages likely to be developed at various points must also be investigated. These decide spacing between lines, and, in some cases, the type of insulating board to be used.

The current to be carried will determine the lower limit of conductor width. It must be remembered that heat developed may affect the bond between the foil and the laminate, and where heavy currents have to be carried—whether from r.f. sources or power supplies—due regard to these must be taken in deciding the width of conductors to be used.

Line width (in)	*Current for 40°C rise in temp.*
0·1	7A
0·05	3·5A
0·03	2·2A
0·01	0·75A

The above table refers to copper foil about 0·003in thick; this is a popular size and covers at 1oz per square foot. For the 2oz weight, the above figures may be doubled. When using this table it must be remembered that a rise in temperature of 40°C is about the maximum that can be tolerated, and a good margin of safety should be allowed.

Separation

For considerations of voltage, a "rule of thumb" is that $\frac{1}{16}$in separation is sufficient for 300V. This applies only to dry air at sea level, and when humid conditions may be met, it will be as well to raise the separation to $\frac{1}{10}$in for 300V. It should also be noted that peak voltages must be the criterion, and an ample margin of safety will prevent breakdowns.

The above will indicate that a considerable degree of miniaturising can be achieved without running into trouble, especially where transistorised circuits are being used.

It is useful to begin a design by setting out the actual components to be used on a piece of card the size of the finished circuit. Conductors can then be put in in pencil to arrive at a tentative layout, and it is a good idea to colour r.f.-carrying conductors in a distinctive way so that the ideal of short leads can be aimed at readily. The need to minimise cross-overs will require a good deal of altering about of the less fixed components, and eventually a reasonably good arrangement will be found. The mathematics of topology is highly complex, and fortunately is not needed in practice, since unavoidable cross-overs can usually be carried out by using necessary components such as resistors and capacitors to act as bridges. It is seldom necessary to use a "jumper"—a piece of insulated wire bridging across a conductor. If more than one or two are used in a design, it usually shows that not enough effort was made to find a better layout.

Screening

It is not necessary to arrange for all the foil to be removed except the conductors. In fact, especially where r.f. circuits are concerned, it is a good idea to let the foil act as a grounded sheet equivalent to the metal chassis of a conventional receiver. However, there are precautions to be taken; since metal will later be removed to isolate the conductors, this may well introduce long return paths for r.f. currents. If each r.f. circuit is laid out so that circulating currents are minimised and, more particularly, localised, much improvement can be obtained in this direction.

"Sensitive" leads may be afforded partial screening by bringing earthed foil close to them on each side, in much the same way as a screened lead is commonly drawn in a theoretical diagram. Also, the earthed foil can be used to effect closure of an i.f. transformer in the same way as does the chassis in normal construction.

When a good layout has been arrived at, the preparation of a full-size drawing can be started. For simple circuits, this may not be thought necessary, but it is worthwhile if only that comparison with the theoretical diagram can be made at leisure. If the technique for preparing the printed circuit is to be merely painting the conductors with a resist ink, the drawing may even be put direct on to the laminated board. If, however, the photographic method to be described shortly is to be used, a good drawing will be needed in any case. It should be done in good black pencil on a sheet of white paper, the exact size of the circuit board to be used.

Where a component is connected to the board, a special termination often called a "land" is provided; it consists of an area of approximately circular shape allowing for a good-sized fillet to secure the component lead to the foil, and should be at least $\frac{1}{10}$in greater in diameter than the component lead. For the usual lead wires of resistors and capacitors a termination $\frac{1}{8}$in

in diameter is enough, but where flexible leads are to be soldered to the foil, at least $\frac{1}{4}$in diameter should be allowed to provide for accidents.

It is unlikely that the home constructor will wish to try dip-soldering, but if he does it is necessary for all components to be mounted on the insulated side of the laminate. This is a design limitation which is normally only necessary for manufacturers, and a more flexible layout is usually possible if hand-soldering is used.

Preparing the Circuit Board

The method of painting the board with a resist ink ready for etching will not be described, as the procedure is obvious. It is only suitable for large circuits in any case, since the width of conductors can only be controlled with difficulty, and for any but simple circuits it represents an incredibly tedious task. Inadvertent errors take a great deal of time and trouble to rectify, and the final product is not generally very elegant. The method to be described is a photographic one, but the reader need not write it off therefore as being beyond his facilities since only the most rudimentary equipment is employed, and the chemicals used are all readily obtainable and are not dangerous. Naturally a little practice will be needed, but laminate is not used up and little but time is consumed.

The outline of the procedure is as follows: the copper foil is covered with a thin layer of resist material; this is insoluble in water but soluble in an organic solvent such as methylated spirit. On top of this is laid a light-sensitive layer which, after exposure to light, is insoluble in water or in methylated spirit, but, where unaffected by light, remains soluble in water. Thus, when a transparent drawing of the circuit is placed over the prepared board and the whole is exposed to light, the dark parts in the drawing shield the light-sensitive layer, while the light parts of the drawing allow light to harden the layer. Development of the exposed plate is done in warm water; this dissolves the unaffected part of the layer and exposes the spirit-soluble resist below, while the resist layer is protected from the action of spirit by the hardened layer above. All that is now needed is to dissolve the exposed resist with spirit, and the board is ready for etching. In fact the whole procedure only takes a very short time to carry out from beginning to end, *once a good drawing has been prepared.*

The materials needed are as follows: 1oz powdered gelatine, $\frac{1}{2}$ dram crystal violet, $\frac{1}{2}$oz ammonium bichromate (potassium bichromate will do but is not quite so effective although easier to obtain). The above can all be obtained at the local chemist's shop. One tin ($\frac{1}{2}$ pint) "Glitseal" polish, obtained at the nearest "do-it-yourself" shop, or 1oz shellac, and a pint of methylated spirit complete the chemical side of the equipment. In addition, it will be necessary to have some detail (tracing) paper or better still Kodatrace or Permatrace paper, black Indian ink and some drawing instruments and brushes.

Now the complete operation is described step by step. First, from the

circuit drawing, a good tracing should be made using drawing instruments and Indian ink. This should be as accurate as possible. Then all the parts *which are to be etched away* are blacked in with Indian ink, leaving conductors clear. The drawing prepared is thus similar to a photographic negative. Care is needed so that all conductors are continuous as required; if a line of ink interrupts one it will have to be pared away with a sharp knife or razor blade very carefully. The finished drawing must be inspected against a light to make sure that all areas intended to be covered are in fact sufficiently opaque, and that no pin-holes exist in the ink coating. The drawing is then put to dry.

Preparation of the Resist

While the ink is drying, the circuit board is prepared. If "Glitseal" is to be used as the resist, a little is poured into a glass tube and thinned with about a quarter its volume of methylated spirit. To the mixture is added a few crystals of the violet dye and the tube shaken until complete solution has taken place. The colour should be a deep violet through which a 60W bulb can barely be seen. The dye is merely to render the resist easily visible on the copper foil. At the same time a gelatine solution is made up by dissolving a bare teaspoonful of powdered gelatine in a fluid ounce of warm water, and a sensitising solution made by dissolving a piece of ammonium bichromate about the size of a pea, or a little bigger, in half a fluid ounce of warm water. When both gelatine and bichromate are thoroughly dissolved, the two solutions are mixed together. The mixture will keep for a few days provided it is kept in a dark-coloured bottle away from heat. After some hours it will gel, but for use this can be overcome by gentle warming in a bowl of warm water. The mixture is best kept still because air bubbles trapped in it are a long time in disappearing, and air bubbles are a great nuisance.

The copper-clad laminate is now scoured carefully with a piece of cotton wool and household scouring powder to which a little detergent has been added. When the laminate is clean enough for use, the water used for rinsing it should form a complete film; no "water-break" must occur. The surface is now dried by means of clean blotting paper or a current of air, and when quite dry is coated *thinly* with a layer of the violet solution, using a brush. The coating will take about an hour to dry, but it is preferable to leave it for a longer period. When quite dry, a little of the warmed gelatine solution is poured on to the surface and spread out evenly with the finger so that a very flat and even layer is obtained. Any excess should be poured off one corner of the board and the whole put in a dust-free place to dry. It is essential to use artificial light—not fluorescent light—during the gelatine-coating process, but ordinary room lighting will not fog the plate in any practical time.

If plain shellac solution is to be used instead of the "Glitseal" polish it should be made up by dissolving a little shellac in some warm methylated

spirit. The mixture should be easily poured; a thick mixture is no good. It should be filtered through a loose plug of cotton wool and crystal violet added to colour it.

Exposure

When the gelatine layer is quite dry—not merely gelled—the laminate is placed copper side upwards on a firm but not hard surface, such as a magazine. The ink drawing is next put on top of it, and a sheet of fairly thick Perspex or glass placed on top to secure a good pressure contact between the drawing and the gelatine. The next process is the exposure to actinic light.

If an ultra-violet "sun" lamp is available, an exposure of about four minutes is sufficient at a distance of a foot. It one has to rely on sunshine it is best to do the exposure in the morning when dust in the atmosphere is less and the content of ultra-violet light is higher. At 10.30 a.m. in the winter an exposure of about an hour—the plate lying flat on the ground—is about right; on a summer morning half this would be sufficient. However, a few trials will doubtless be needed since all readers do not live in the country where the air is clear and exposures may well have to be lengthened in some places. If on development, it is found that exposure has been insufficient, the laminate may be cleaned off with a small scrubbing brush and hot water, re-scoured to clean up and re-coated with lacquer and gelatine for a further effort. As an alternative, a step-wise test exposure, in the manner of a test strip in photography, can settle the exposure question quite quickly.

Development

Development consists first of washing off the gelatine which has not been hardened by light. This is done in water at 40°C—this temperature feels hot but not uncomfortably so to the hand. Two to three minutes will be sufficient. The plate is now allowed to dry thoroughly once more. When completely dry, the second stage is carried out by pouring on to the surface a small pool of methylated spirit and spreading it out until it covers the plate. Gentle rubbing with a small plug of cotton wool will now remove the lacquer from the places not protected by the hardened gelatine. A second and third wash will be needed, using clean methylated spirit, to remove all traces from the copper foil. At this stage the pattern of the printed circuit will be easily visible.

It is probable that at the first attempt there will be some imperfections in the pattern. If not too bad or too many the plate may be used after re-touching with some of the lacquer on a fine brush. If, however, the blemishes are too extensive or too many, re-touching can be a tedious business, and it will probably be found quicker to clean off the laminate and start again.

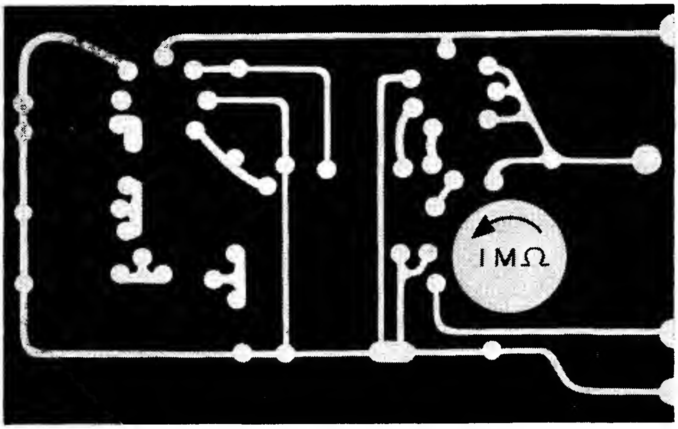


Fig. 1. An example of a printed circuit prepared by the photo-resist process.

Faults Encountered

With only a little practice excellent results can be achieved. Principal faults include the following:

(a) After water-development, the circuit pattern is visible but spirit-development fails easily to reveal the bare copper pattern. This is due to over-exposure.

(b) The lacquer film comes off in places when water-developing the gelatine. This is due to insufficient cleaning before coating.

(c) The image is much more readily developed in some parts of the plate than in others. This is due to too thick a layer either of gelatine or lacquer.

(d) Water development reveals the image but subsequent spirit development takes off all the lacquer coating. This may be due either to under-exposure to light or to the use of too old a gelatine solution.

(e) Water-development reveals the image but when spirit is applied, the image is marred by blotches where the film should have been but has washed off. This is due to uneven coating with gelatine or to under-exposure, or to too hard rubbing with the cotton wool.

When a good image has been obtained, the etching process is undertaken. Experiments show that the fastest etch in ferric chloride solution is obtained when the solution contains 30% ferric chloride. As bought, the ferric chloride is in its hydrated form which increases the weight required to make a solution of a given strength. If four ounces of ferric chloride is obtained—a quite convenient amount—it will need to be made up to a solution of volume 16oz and the water should be adjusted after adding one tablespoonful of concentrated spirit of salt—hydrochloric acid. In this solution a complete etch of foil 0·003in thick is accomplished in 12 minutes if the dish containing the fluid is rocked continuously once per second. If stirring is violent the

etch will be speeded up. It is not advisable to leave the dish to stand or else the etch may well take an hour or more; clearly the less time taken, the less chance there is of the lacquer coating failing.

After the etching process is complete, the board is removed from the solution and washed thoroughly under running water for a few minutes, after which the lacquer may be removed by scrubbing under hot water in case some gelatine has survived the etch bath, followed by a spirit wash and final cleaning up with some scouring powder and water. When dry, assembly of the circuit can commence.

Alternative Procedure

When large areas of copper are to be etched away, correspondingly large areas need to be blacked out in the drawing. The application of a sufficiently dense and even coating of Indian ink over large areas is a difficult matter and, if ordinary tracing paper is used, the problem of cockling arises. In such cases it will probably be best to use such a paper as Permatrace and to use poster-paint instead of Indian ink. Poster-paint, however, tends to crack if the paper is flexed too much, so care will have to be taken to keep it reasonably flat. As an alternative, some success has been had by depositing dyed gelatine direct on the foil and, after exposure under a "positive" drawing (conductors in black) and water-development in the usual way, lacquering in the conductors direct—using the gelatine as a stencil. Hot washing-soda solution will remove the hardened gelatine leaving the resist-covered conductors behind. However, plastic laminates do not stand up too well to alkaline solutions, and a trial run with some pieces of scrap laminate is advised as a preliminary.

Figure 1 on page 173 shows a simple circuit board produced by the two-layer method described in previous paragraphs. Little re-touching of plate or finished circuit has been done; most defects seen are due to imperfections in the drawing. That there are imperfections is clear, but it should be noted that this was a deliberately-chosen difficult circuit because of the large areas to be etched away. More normal circuits would present an even better appearance with much less difficulty.

Chapter Twelve

A TV BABY ALARM

SOME experimenters may have already tried connecting a microphone to the audio stages of a television receiver in an endeavour to superimpose the sound of distress in an infant's bedroom upon the programme sound. Direct connection of a microphone is rarely successful owing to the relatively small gain of the audio section on the one hand and the low level microphone signal on the other. This article describes how the microphone signal can be boosted before it is fed to the audio section.

Triode-Pentode Audio Section

The majority of television sets employ a triode-pentode audio circuit as shown in Fig. 1. Here the signal from across the load of the sound detector is applied to the grid of the triode and the triode steps up the audio to a level suitable for driving the pentode output stage and loudspeaker. The triode employs "grid-current" biasing by the 10MΩ resistor between grid and chassis and the pentode has normal cathode biasing by the 330Ω cathode resistor.

Slightly different arrangements will be found between sets of different make and model, but all models using a triode-pentode valve will have a circuit almost exactly as shown in Fig. 1, plus, possibly, some form of negative feedback, depending upon the type of valve used. The PCL82 has separate cathodes, but some triode-pentodes have just a single cathode common to both sections and a slightly different circuit, of course.

For average sound volume the triode receives at its grid about 50–100mV of audio signal. The sound detector produces a signal across its load several times stronger than this, but this high level is used only when the volume control is turned full up, and this is not very often, provided the set is working correctly and the viewers have average hearing.

Now, for a baby alarm to be fully effective, even the slightest murmur from the infant should be audible over the programme sound, irrespective of the setting of the volume control. This means, then, that the alarm channel must always be at full gain.

This presents several problems such as signal-to-noise ratio and hum level, bearing in mind that it is neither desirable nor possible to position the microphone close to the face of the baby. Of course it would be a bad thing for the alarm channel to superimpose hum and noise (other than that of the infant) upon the sound of the programme.

Transistor Microphone Amplifier

The solution to the problem was discovered in a small transistor microphone amplifier, as shown in Fig. 2. This is a simple common emitter circuit with the microphone signal applied to the base via a microphone transformer. The amplifier is powered from the h.t. line of the receiver via a potential-divider network comprising R1 and R2. With an h.t. line of about 230V, approximately 10V d.c. occurs across R2. Although this voltage has already been smoothed by the receiver smoothing circuits, extra smoothing is given by a 1000μF electrolytic capacitor C1. This ensures that the amplifier does not introduce extra hum into the grid circuit of the audio section.

As the amplifier will often be coupled to an a.c./d.c. type of set where the chassis is connected to one side of the mains supply, extra care must be taken to isolate the microphone feed circuit. This design incorporates features that render it absolutely safe in use.

Safety First

Firstly, isolation is given by the A-B and C-D windings (primary and secondary) of the microphone transformer T1. Secondly, isolation is provided by capacitors C2 and C3. Two series-connected capacitors are used so that in the remote event of one capacitor breaking down there will still be another in the circuit maintaining adequate safety. As an added precaution, the outer screen of the microphone cable can be connected to a true earth, such as a rising water main; but make sure that a *true* earth is used.

The amplifier is designed for a moving coil microphone of about 20Ω to 100Ω impedance (the actual value is not important), and it should be connected to the primary of the microphone transformer through a thin, coaxial cable. The type of cable used as television aerial downlead in areas of high signal is ideal and not very costly. This cable can be obtained with a cream plastic covering, and if it is necessary to run it to the bedroom on the surface of woodwork, such cable is far better than that with the more conventional brown plastic covering.

The amplified signal is developed across the collector load resistor R4, and is fed to the audio stages of the set through the coupling capacitor C5 and the "buffer" resistor R5.

Connections to the Set

There are three points of connection from the set to the amplifier, marked 1, 2 and 3 in Figs. 1 and 2. Reverting to Fig. 1, it will be seen that a connection is made from the chassis on 1, from the slider of the volume control on 2 and from the h.t. line on 3.

The h.t. connection in particular must be well insulated. The connection to the volume control should be through screened cable, and a length of the

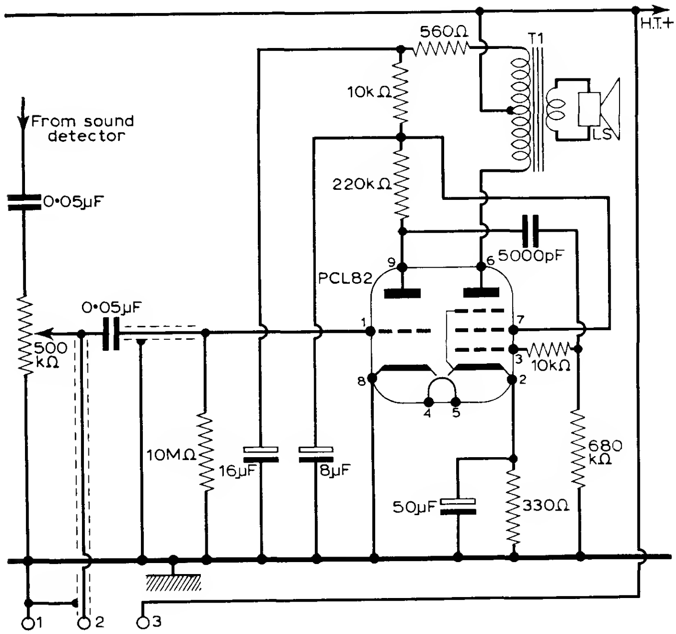


Fig. 1. Circuit of a typical television sound-output stage, using a PCL82 triode-pentode valve. The connections terminated at 1, 2 and 3 are extra to the circuit and used for connecting the amplifier of the baby alarm.

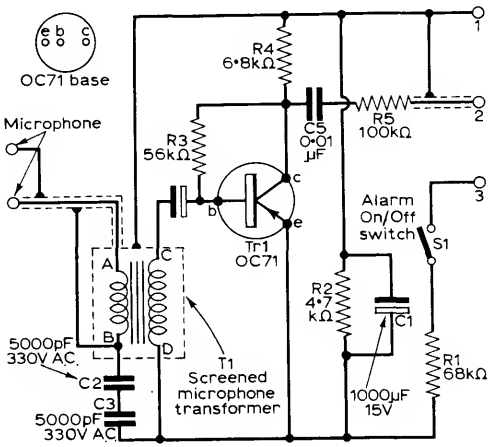


Fig. 2. The circuit of the amplifier. The isolating capacitors C2 and C3 must be no larger in value than 5000pF (5nF) and must be rated at 330V a.c. at least.

coaxial microphone lead can be used here quite successfully. It will be understood, of course, that the amplified microphone signal is applied direct to the grid of the audio triode, meaning that full microphone-channel gain will always exist provided the volume control is not turned right to zero, for then the grid is shunted by the 0·05μF coupling capacitor. While this will not reduce the gain completely it will considerably reduce the alarm signal. If required, this can be overcome simply by connecting a 47kΩ resistor between the slider of the volume control and the 0·05μF capacitor. Such a resistor would not normally affect the operation of the volume control on the programme sound, and in some sets it may already be fitted.

Amplifier Construction

The transistor microphone amplifier is best made up on a piece of $\frac{1}{16}$in laminated plastic board measuring about 2in × 5in, as shown in Fig. 3. The components may then be secured to the board by threading their lead-out wires through small holes drilled with a No. 55 or similar drill.

The wiring may then be completed on the reverse side of the board, and the broken-lines in Fig. 3 show this. Note that where wires cross, a length of insulated sleeving must be threaded on to the conductors to avoid short circuiting, which could easily damage the transistor.

The screened cables on the microphone circuit and the one from point 2 to the set should be secured to the board by small soldering tags, the latter retained by 6 B.A. nuts and bolts.

After the wiring is completed and the circuit has been checked several times, the amplifier can be mounted on the inside of the cabinet (if of wood) by drilling a hole in each corner of the board and using wood-screws and stand-off pillars to prevent the circuit from touching the cabinet—a practice which could incite pick-up of hum.

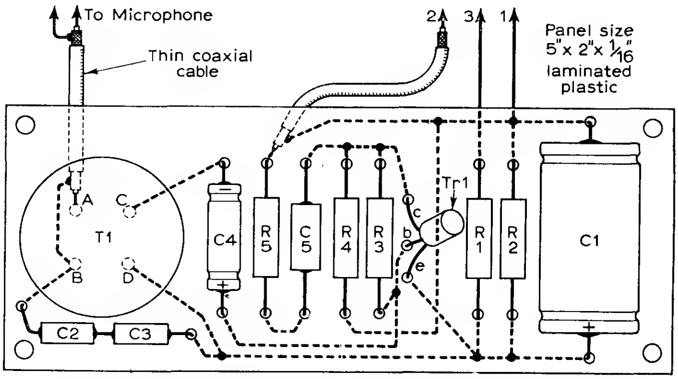


Fig. 3. The amplifier can be built on a plastic panel measuring about 2in × 5in. The components are all mounted on one side of the panel.

If it is not possible to fix the board to the cabinet, then it must be secured by stand-off insulators to the chassis or to a chassis member, ensuring that the parts are kept well clear of the chassis and components. The amplifier should also be kept as far as possible away from the a.c. and power circuits of the set, including the field and line output stages, where the large signals could get into the amplifier and superimpose a hum or whistle on the sound.

As it will not be desirable to keep the amplifier on at all times, a simple well insulated on/off toggle switch should be interposed between the lead from R1 to the h.t. positive line. This switch can be fitted to the rear cover of the set, and should be of the all-insulated type which has no metal exposed at all.

Components

All the components for the microphone amplifier are readily available from most radio stores. Radiospares Ltd. markets a suitable microphone transformer and 1000μF capacitor for C1.

One of the ex-government or surplus moving coil microphone inserts can be boxed in such a way as to fit in with the nursery furnishings. Such a microphone is usually reasonably sensitive for this application; but other microphones and inserts all ready and made up can be used if required.

A high impedance ceramic or crystal microphone could be used (without the transformer T1) but this is not particularly recommended owing to the noise and hum picked up on a high impedance line from the amplifier to the microphone. An ordinary coaxial cable tends to become microphonic under high gain conditions, and although on the face of it it may seem less costly to use a crystal microphone and no transformer, in practice this is not always the case after one has found it necessary to buy special cable to clear hum and noise. If a high impedance microphone is used, however, *mains isolation is still essential.*

Operation

The device is best checked by removing the aerial from the set (or tuning to a vacant channel), turning the volume control halfway on and checking how well noises from the baby's room are reproduced in the set. At this stage the microphone should be adjusted for the best position by using a ticking alarm clock as a substitute for the infant.

Signal may then be applied to the set and the alarm break-through ratio assessed. In most cases the alarm signal will always be well above the programme signal.

Safety

As a further safeguard the TV set should be wired to the mains in a three-pin plug in such a manner that the chassis is connected to the neutral side of

the supply, not the live side. This can easily be checked with a neon test-screwdriver.

List of Components

Resistors:
- R1 68kΩ1W
- R2 4·7kΩ
- R3 56kΩ
- R4 6·8kΩ
- R5 100kΩ

Capacitors:
- C1 1,000μF electrolytic 15V
- C2 5,000pF 330V a.c.
- C3 5,000pF 330V a.c.
- T1 Microphone transformer (Radio-spares)
- Tr1 OC71 transistor
- S1 s.p.s.t. toggle switch (all-insulated type)

Chapter Thirteen

THE SUPPRESSION OF INTERFERENCE

THE best way of proving whether or not an electric tool or other domestic electrical appliance is a source of radio or television interference is to set up the appliance as it would normally be used and observe the effect on one's own receiver. Before the appliance is switched on, however, the receiver should be tried to ensure that it is free from interference and then it will be conclusively evident that any interference from the loudspeaker or on the screen is caused entirely by the appliance under test.

Wear and Vibration

Almost all recently made tools and electric motors are suppressed for both radio and television, but, even with this type of equipment, there is always the possibility that wear on the motor brushes or commutator will eventually cause interference. Continuous vibration may also loosen the internal suppression components causing random bursts of very heavy interference when the appliance is in use.

What is Reasonable Suppression?

It should be understood that perfect interference suppression is virtually impossible to achieve. This would necessitate costly anti-interference precautions at both the receiver and source of interference. Interference precautions at the source are, therefore, based on the assumption that the receiver itself is installed with due regard to interference pick-up. For example, if "reasonable" suppression is undertaken on the appliance and yet a near-by neighbour suffers interference because his aerial is mounted, say, in the roof-space close to the mains wiring, then it becomes the responsibility of the neighbour—viewer or listener—to improve his installation before making a complaint to the Post Office. A radio dealer would advise on such a problem and, if necessary, the dealer himself may consult the Post Office.

The Effect of Signal Strength

How much interference is seen on a picture or heard from a loudspeaker depends not only on how much interference the appliance is radiating but

also on how much signal the set is receiving. If the wanted signal is weak, the interference may be equal to or above the wanted signal, in which case it would override the picture and sound and cause extremely disturbing noises and flashes. On the other hand, if the wanted signal is much stronger than the interference, then it may be almost totally masked by the wanted signal and hardly show up at all.

Clearly, then, in areas where the signals are strong (close to a transmitter), a far greater level of radiated interference can be tolerated than in areas where the wanted signals are weak. This means that "reasonable" suppression in a strong-signal area can be "poor" suppression in a weak-signal area.

Manufacturers of electrical appliances usually design their suppression circuits on the assumption that there will be an average signal level at the receiver, of a level, for instance, that would work a television set adequately from an H-type aerial. Such suppression would, therefore, be almost perfect in areas close to a powerful station, but may be less than "reasonable" in areas where the signal strength is very weak.

Radiation from Appliance

Interference to television reception is almost entirely radiated direct from the appliance and connecting cable. The appliance and cable, in fact, act rather like a transmitting aerial and, the closer the appliance to the receiver, the stronger will be the interference. Thus, when an appliance is checked on one's own receiver some small interference effect will almost certainly be observed. This test should, of course, be made with the set receiving a signal, for in the event of there being no programme, the receiver would be highly sensitive to even the smallest amount of interference.

If the interference effect appears to be excessive on the near-by receiver, it often pays to visit neighbours to see whether their reception is being affected. If the interference is highly disturbing, steps should be taken to improve the suppression of the appliance.

Cause of Interference

Interference of the nature under discussion is caused entirely by sparking taking place within the appliance. Whenever there is a make or break in an electrical circuit, sparking takes place. This happens even when an ordinary electric light switch is operated and may cause a small click on sound and an instantaneous flash on vision. Such interference is so short-lived that it is not troublesome. However, when sparking is prolonged the resulting interference can be very disturbing.

Prolonged Sparking

Prolonged sparking takes place between the brushes and the commutator of electric motors having armatures as shown in Fig. 1. As the armature

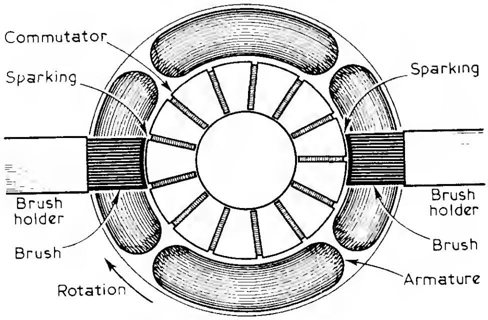


Fig. 1. Interference can be caused by sparking between the brushes and commutator of electric motors. The trouble is aggravated by wear of brushes or commutator. Induction motors (which have no brushes) do not give rise to interference.

and commutator rotate, so the current is disconnected and connected very rapidly and the sparking is almost continuous. This produces a continuous buzz on sound, and bright horizontal lines across the picture. The amount of interference radiated is almost proportional to the size of the spark, which means that on a motor with worn brushes or commutator the interference level is likely to be very high indeed.

Similar sparking occurs between the contacts of a thermostat, as shown in Fig. 2. This happens only when the control comes into or goes out of operation. With a faulty thermostat, however, the sparking on disconnection can be prolonged due to arcing between the contacts, so instead of short-lived interference, prolonged buzzing and flashing may well be produced. Electric irons and electric blankets cause this type of interference, but immersion heater thermostats are often sufficiently well screened by the earthed water-jacket which surrounds them, although they also may cause interference.

The Cure

If the appliance is obviously defective in itself, then this should be attended to before suppression or additional suppression is contemplated. It should be noted that there are two basic suppression arrangements—one for radio and the other for television—and it does not necessarily follow that radio suppression will have been achieved after dealing with an appliance for television, and vice versa. Where an appliance is affecting both radio and television sets, it is necessary to combine the suppression arrangements recommended for each.

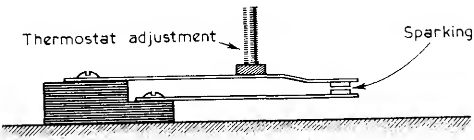


Fig. 2. Interference can be caused by sparking between the contacts of thermostats. The arcing may be prolonged on switch-off if the springs of the thermostat are weak.

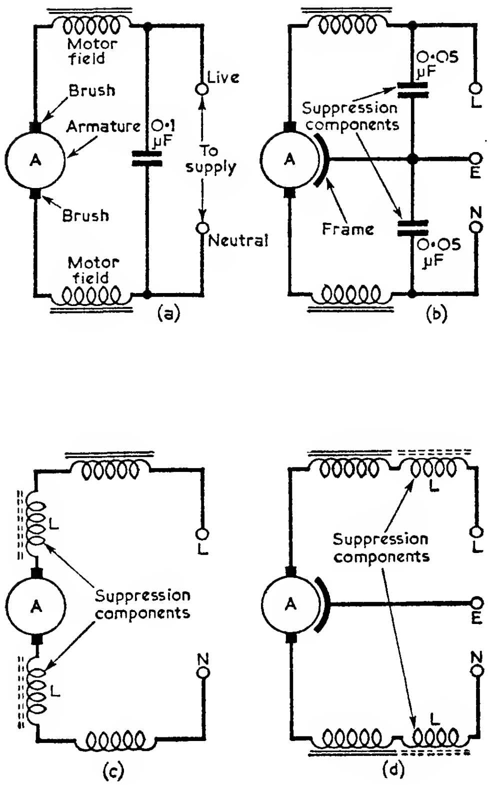


Fig. 3. These suppression-circuits are designed for use on small appliances in areas of reasonably good signal strength and moderate interference: (a) and (b) are for radio and (c) and (d) for television.

Suppressing TV Interference

When suppressing a motor for television interference, it is essential that the suppression components be mounted as close as possible to the part of the motor concerned. For example, if a choke is to be connected to the brushes, then it *must* be mounted inside the motor housing adjacent to the terminal of the brush holder. This does not follow to such an extent with radio suppression, for such interference is mostly mains-borne, and the principle in this case is to prevent the interfering currents from travelling out of the appliance into the mains wiring. We have already seen that with television interference, direct radiation is the important factor.

Suppressing Radio Interference

Small chokes and capacitors have been developed by various firms specialising in suppression equipment, for inclusion inside the motor housing, but, even so, ingenuity is often necessary in order to accommodate a complex

suppression arrangement inside a portable appliance. If radio as well as television suppression is required, it may be necessary to fix the radio suppression components outside the housing.

With this in mind, Belling & Lee have designed a series of suppression units that can quickly be fitted to the flex lead within about 12in of the appliance. Similar arrangements in the form of adaptors for plugging in at the power output sockets are also available from radio dealers. The latter type are ineffective on television, but the flex-lead type are also suitable for television when moderate suppression is required provided they are connected close up to the offending appliance.

Moderate Suppression

In Fig. 3 are given four circuits for moderate suppression of portable appliances. At a is the simplest of radio suppression circuits and takes the form of a capacitor across the mains input terminals. If the appliance uses three-core cable (e.g., with a green earth wire), the arrangement at b should be used. Special capacitors for this purpose are available.

Similarly, at c is shown the simplest of television suppression circuits which uses a pair of television chokes connected close to the brushes. At d is the arrangement for three-cored cable, but here it may be better to include

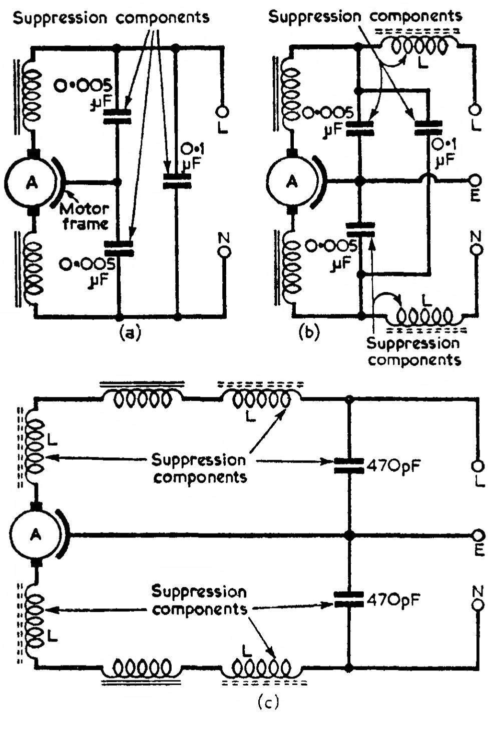


Fig. 4. These circuits are also used for small appliances but give better results than those of Fig. 3. Circuits (a) and (b) are for radio, and (c) is for television.

the chokes L as in c, depending on results. When checking the success of suppression, it is essential to reassemble the motor housing and run the unit in the normal way with all the screws properly tightened.

Other Circuits

In weak-signal areas, suppression circuits after the style of those in Fig. 4a, b and c are often called for. Fig. 4a is for radio and could be used on a two-terminal or three-terminal appliance, but if only two conductors are used, it is still necessary to take the junction of the two $0{\cdot}005\mu F$ capacitors to the motor frame. Connection should be made to a soldering-tag clamped beneath one of the motor's securing screws.

The circuit at b is also for radio and gives even better performance than a. This is due mainly to the two r.f. chokes L, which in this circuit should have an inductance of about 1·5mH.

The arrangement at c is for television only, and the improved performance is given by the use of four television chokes and the two 470pF capacitors. As already pointed out, it is essential that all these components are mounted as close as possible to the motor unit, preferably in the housing.

Suppression Components

For television suppression of small appliances there are two basic components, small chokes wound on dust-iron cores and small 470pF ceramic capacitors. The chokes are rated at 1A, 2A or 3A, according to the current requirements of the appliances, and the capacitors must be rated at 300V a.c., at least.

Capacitors for radio suppression should have a rating of 250V a.c. and delta-connected units containing three capacitors, as in Fig. 4a and b are available in values of $0{\cdot}1\mu F + 0{\cdot}005\mu F + 0{\cdot}005\mu F$ and $0{\cdot}1\mu F + 0{\cdot}01\mu F + 0{\cdot}01\mu F$. Chokes for radio suppression are somewhat larger than their television counterparts, depending upon inductance and power rating, but are also available in 1A, 2A and 3A ratings.

Fitting Suppressors

When fitting, extreme care should be taken to avoid the possibility of internal short-circuits and adequate insulation should be adopted throughout. The components should be mechanically rigid and wired in a sound electrical manner. Capacitors greater than $0{\cdot}05\mu F$ should never be connected to the frame of an unearthed appliance, but this should not be taken to mean that it is safe to run an appliance without an earth. All appliances featuring an earth conductor *must, for safety reasons, be connected to an efficient earth point*, such as to the earth pin of a three-pin plug.

Chapter Fourteen

A DEAF-AID FOR TV

THOSE who unfortunately have impaired hearing often have difficulty in enjoying a television programme and become singularly unpopular by turning up the sound by an extra 6dB or so. An ordinary deaf-aid is not at all successful for listening to television or radio sound, since the microphone of the deaf-aid picks up not only the sound from the loudspeaker of the set but also sound reflected from the walls and ceiling of the viewing room and this, along with acoustic coloration endowed by the characteristics of the room itself, results in the sound from the deaf-aid earpiece being very much confused.

Slightly better results are possible by locating the deaf-aid microphone as near as possible to the loudspeaker, but with television this usually means viewing the picture far too closely, which could cause eye trouble.

To assist keen viewers whose auditory responses are somewhat below normal, a series of experiments was carried out on television sound channels, and in this chapter is explained the best, and most inexpensive, way of solving the problem for at least 90% of hard-of-hearing viewers.

Isolation

All along, the aim was to achieve total isolation between the receiver and the earpiece, for modern receivers represent a potential source of danger to users of devices and circuits connected to the inside of the set. This is because the metal chassis of the receiver is invariably connected *direct* to one side of the mains supply, and if this happens to be the "live" side a person making simultaneous connection with this and an earthed object would really be in trouble—aggravated, of course, by the "live" connection being made to the head (via an earpiece or associated wiring).

However, if special care is given to isolation, the danger outlined above is eliminated. The rest of this chapter, therefore, describes a deaf-aid system which requires a direct electrical connection to the receiver, but which is perfectly safe and isolated from the mains supply in more than one way.

Circuit

The circuit of the deaf-aid is shown in Fig. 1. Transformer T1 serves two purposes, it steps up the voltage at the speech coil of the internal loud-speaker to a suitable value for operating the earpiece after allowing for the

attenuation of the tone control, and furthermore it isolates the receiver from the earpiece. Further isolation is given by capacitors C1 and C2.

VR2 is the ordinary manual volume control, and VR1 is a pre-set volume control which can be adjusted when the parent receiver is set for normal volume and with the manual control at mid-range. The tone control is continuously variable and really acts as a treble-cut control in conjunction with C3. This is extremely useful for securing optimum intelligibility, as the auditory frequency response varies somewhat between persons with impaired hearing and the tone control allows for individual setting of the response.

Connection to Parent Receiver

Fig. 2 shows the basic circuit of the output stage of most television sound channels. The primary of the loudspeaker transformer is connected between the anode of the output valve and the h.t. line while the secondary is connected across the speech coil of the loudspeaker.

On some receivers the speech coil is connected at one side to the chassis of the receiver. This may or may not be important, and to achieve optimum mains isolation for our purpose it is best if such a connection can be eliminated. This can be done without ill effect provided the other tag of the speech coil or transformer secondary does not go back into the set as a negative feedback loop. If it does, then the connection must remain, otherwise instability and sound distortion may result.

Fig. 3 shows the arrangement in pictorial form. The loudspeaker transformer may not be mounted so accessibly on the chassis as the illustration suggests. Nevertheless, the connection to its secondary can easily be followed from the loudspeaker tags, working backwards, and negative feedback wires or wires to the chassis will be identified at the same time.

Fig. 4 shows how the complete unit can be mounted in a small electrical box. The box used by the author is List No. 2031 by MK Electric Limited and the matching frontplate is List No. 3830, both readily available from electrical shops.

The earpiece—which should be of medium-impedance or high-impedance—is connected through two sockets, while the lead to the set actually emerges through a hole in the side of the box. The reason for the latter is to ensure that there are no exposed pieces of metal—such as plugs and sockets—which, in the event of a fault in the parent receiver (or a chassis-connected secondary winding on the loudspeaker transformer) could become "live" and give an electric shock to the user.

There is, of course, no danger whatever of the wearer of the earpiece ever getting a shock since it is isolated from the set in two ways—by T1 and by C1 and C2. A good quality, ivory (or other colour as may be required) plastic-covered two-conductor cable should be used between the unit and the set, and then the unit can rest on a table or on the arm of the chair of the deaf viewer, who may then conveniently adjust both volume and tone as required.

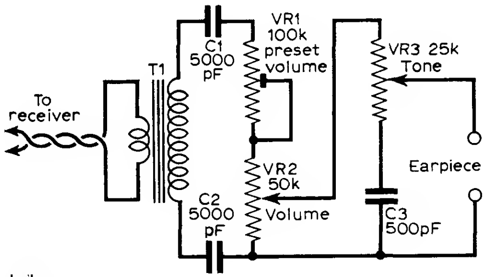


Fig. 1 (right). *The circuit of the deaf-aid. Note that the low-impedance winding* (*secondary*) *of T1 is connected to the receiver.*

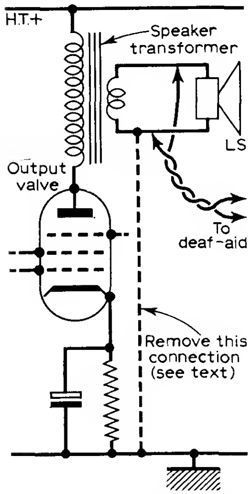
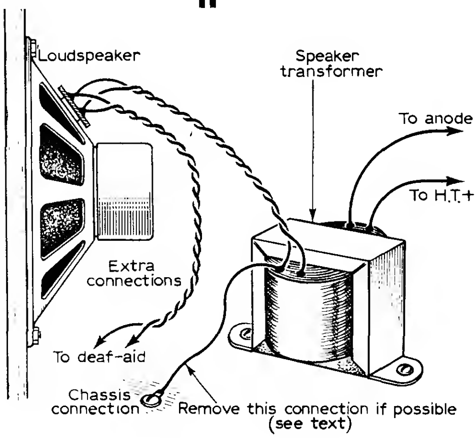


Fig. 2 (above). *Connecting the deaf-aid to the receiver.*



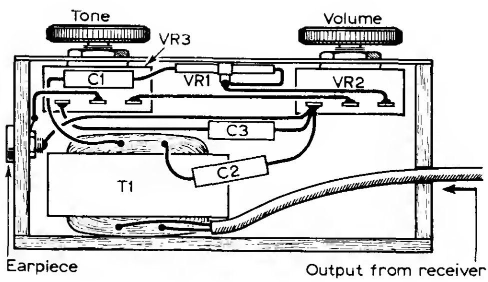


Fig. 3 (above). *Pictorial representation of the connections to the receiver.*

Fig. 4 (left). *A suitable method of constructing the deaf-aid.*

Setting Up

The unit should be set-up in the following manner. After connection to the parent receiver, the receiver's volume control should be adjusted during the course of a programme for the normal listening volume as required by the viewers with normal hearing.

The person needing the aid should then set his manual control to mid-range and adjust the pre-set volume control, in conjunction with the tone control, for the most comfortable listening level. For safety, insulated tools should be used when making these adjustments.

The top may then be put back on the unit and the main volume and the tone controls used normally during a programme. Alterations to the settings of the deaf-aid controls will not, of course, affect the volume of sound delivered by the loudspeaker in the receiver; but an alteration of volume of the set will call for a corresponding adjustment of the deaf-aid's volume control. See also paragraph on *Safety* at the end of Chapter 12 on page 179.

LIST OF COMPONENTS

Component	*Make*
Ivory or brown plastic box, List 2031	MK
Ivory or brown frontplate, List 3830	MK
Midget loudspeaker transformer (T1)	Radiospares
Two midget linear volume controls (VR2 and VR3)	Radiospares
Two single insulated sockets (for the earpiece)	Radiospares
100kΩ slider pre-set (VR1)	Radiospares
Hi-K midget ceramics (C1, C2 and C3)	Radiospares
Wire, etc.	
Earpiece	

Chapter Fifteen

NUVISTOR CIRCUITRY

RIGHT from the beginning it has been the aim of electronic engineers to create equipment for the amplification of very weak radio signals which itself does not add spurious noise signals to the radio signals. The maximum usable sensitivity of any radio or television amplifier is governed directly by the amount of noise signal that the process of initial amplification produces.

High-gain amplifiers are relatively easy to make, but these are of very little use if the signal to be amplified is so weak that the noise generated masks the amplified signal.

A television picture free from noise is only possible when the noise contributed by the initial amplification is at least 200 times (46dB) down on the required signal. This makes it clear that the weaker the signal to be amplified, the better must be the noise performance of the amplifier itself to maintain the ratio of 200:1.

Noise is generated not only in the amplifier but also in the aerial, and then, of course, there is also noises attributable to general static and space signals from stars and so on. Excluding the first amplifiers in the receiver, therefore, some noise will always be present resulting from the mode of radio propagation as we know it at present.

Nevertheless, under practical conditions, the majority of the noise is produced in the first amplifier, for after that, the signal is usually strong enough to outweigh the noise contribution of subsequent stages.

Much work has been undertaken to enhance the noise performance of low-level amplifiers, and in this sphere the transistor is being found to be of considerable help, because here there is no "thermal" noise resulting from the emission of electrons from a hot cathode to a positively charged anode.

The Nuvistor

Competition to the transistor in this respect was offered by the recent Nuvistor valve. This is either a triode or tetrode valve of special design, which in the past was of American origin, under the designation 6CW4. However, a Mullard counterpart has been evolved under three main versions. The 7586 which is a medium-μ triode, the 7895 which is a high-μ triode and the 7587 which is a sharp cut-off tetrode.

Although Nuvistors are really designed for professional and industrial

applications, they may, nevertheless, prove of interest to the experimenter. Indeed, one excellent design for a Nuvistor Band III Amplifier is described in Chapter 1 on page 11. This employs the American 6CW4, but there is no reason at all why the Mullard high-μ version could not be used instead.

What is a Nuvistor?

The type of electrode structure employed in the Mullard series of Nuvistor valves is based on a concentric arrangement of cylindrical electrodes. These are supported by three pins which project through a ceramic base plate (see Fig. 1). The valve is finally encased in a metal shell which needs to be adequately bonded to the chassis of the amplifier for optimum stability.

Although pins are available for earthing the metal shell it is rather important that something better in the way of earthing is produced by means of the earthing lugs on the metal shell. During the course of experimenting with the American version, it was found on several occasions that instability tendencies resulted from poor r.f. earthing of the shell, even though the earthing pins were in excellent d.c. contact with the chassis.

Characteristics

The valves can be used with either grid-leak (e.g., grid current) biasing or conventional cathode biasing. The low internal capacitances coupled with the very small lead inductances permit the valves to be used in the earthed-cathode mode, with the input signal applied to the grid. Under this condition, however, neutralisation is necessary to secure optimum stability and noise factor.

Basic Circuits

In Fig. 2 is shown a basic amplifier circuit using grid-leak biasing and capacitive neutralisation. Here it will be seen that the cathode is strapped direct to chassis and that a resistor is used in the grid circuit. Owing to activity between the cathode and grid, a small potential develops across the grid resistor, and it is this which is used to bias the valve.

This type of biasing usually permits a slightly greater gain to be obtained from an r.f. stage as compared with cathode biasing. The noise figure is also slightly better with grid-leak biasing in most cases, depending upon how well the unit is designed, neutralised and mechanically constructed.

Neutralisation is effected in Fig. 2 by the pre-set trimmer C1, and if the amplifier is to tune over a band of frequencies, it is best to adjust for optimum neutralisation at the low-frequency end of the band. This does not apply, of course, to v.h.f. radio and television pre-amplifiers, for then it is the usual practice to adjust at the vision-carrier frequency.

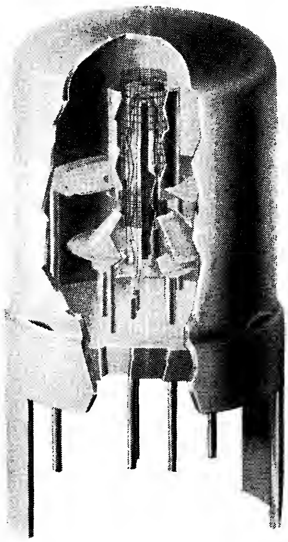
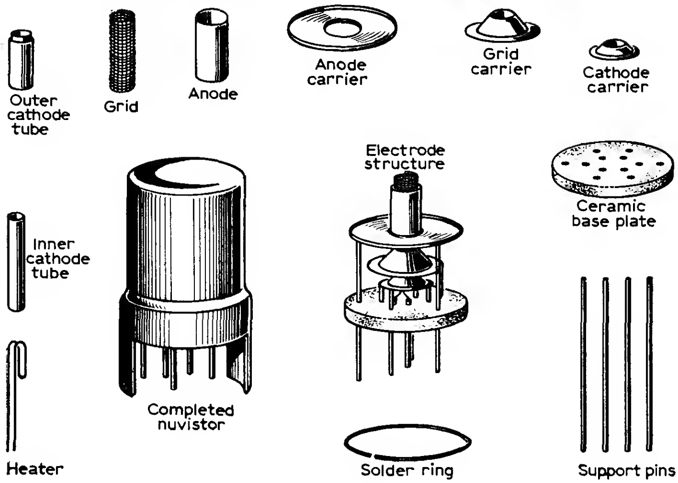


Fig. 1. The Mullard Nuvistor Triode.

left—*a general view*

right—*a cut-away view showing the interior*

below—*the component parts used*



Neutralisation Adjustment

Possibly the best way that the experimenter can adjust the circuit for optimum neutralisation is first to disconnect the l.t. feed to the valve, then apply a very strong signal at the required frequency and finally adjust the neutralising trimmer for minimum output.

This is fairly easy to undertake when the stage takes the form of a pre-amplifier in front of a television receiver. The applied signal can either be

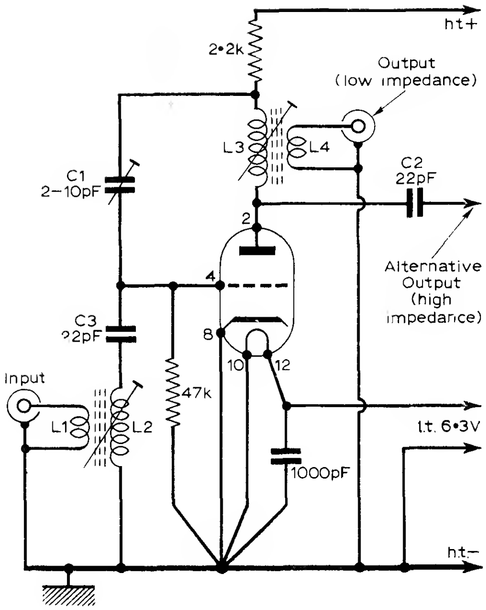


Fig. 2. The circuit of a v.h.f. amplifier using a Nuvistor triode. This is arranged in the earthed-cathode mode with grid-leak bias and capacitive neutralisation. Note the common earthing-point on the chassis.

on the sound or vision-carrier frequency. If the former, the receiver's volume control should be set at maximum and sufficient input signal applied at the aerial terminal of the pre-amplifier to be heard adequately in the loudspeaker (the signal being modulated, of course).

The neutralising trimmer is then adjusted for minimum output at the loudspeaker. On the vision carrier, a modulated signal will give horizontal bar patterns on the raster, and the neutralising adjustment should be carefully set to minimise these.

Alternatively, the adjustment can be made for the best noise performance, but as this requires a noise generator it can rarely be undertaken by the experimenter and, in any case, there is very little difference in the overall performance whichever method is used.

The circuit of Fig. 2 shows a common earthing point for the various components, and this should be maintained so far as any additional decoupling of r.f. by-pass components are concerned. The base connections and dimensions of the valve are given in Fig. 3.

The circuit shown would be suitable for a television pre-amplifier on any channel in Bands I and III. As a television pre-amplifier, low impedance coupling windings would be used to apply the aerial signals and extract the amplified signal—L1 and L4 respectively. If a high impedance output were required, however, L4 would not be used and the signal would be extracted direct from the anode, via C2.

The number of turns and mode of construction for the coils L2 and L3 will depend on the channel which it is required to amplify. Normal coil-winding techniques should be followed, and the dust-iron cores should be used to provide a range of inductance control for tuning. The low impedance coupling coils should be positioned towards the C3-end of L2 and the C1-end of L3.

Inductive Neutralisation

A slightly modified circuit is shown in Fig. 4, where inductive neutralisation is used instead of a capacitor and cathode bias is used instead of leaky-grid bias. R1 provides the bias due to the voltage drop across it, while r.f. across the resistor is by-passed by C3. The grid of the valve is returned direct to chassis through L2, and neutralisation is effected by L5. C1 here acts solely as a d.c. blocking capacitor to prevent h.t. from reaching the grid circuit and being shorted through L2.

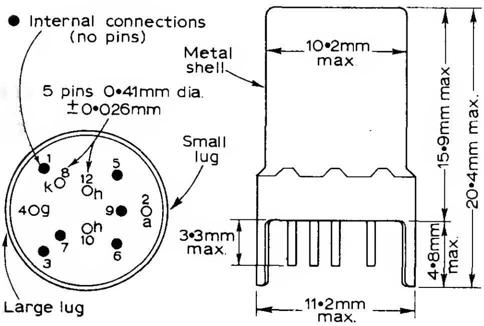


Fig. 3. The base connections and dimensions of the Nuvistor.

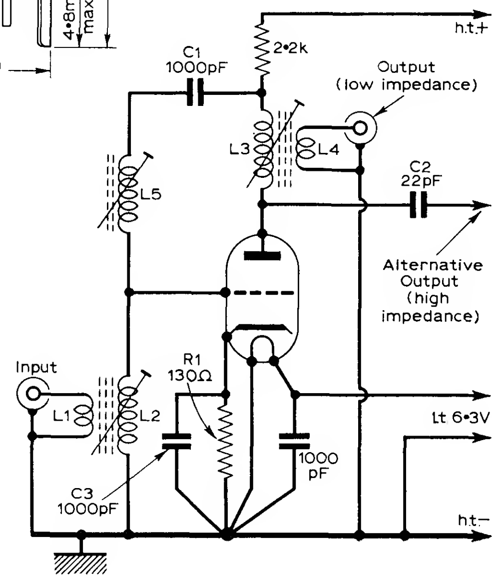


Fig. 4. A Nuvistor circuit using cathode bias and inductive neutralising.

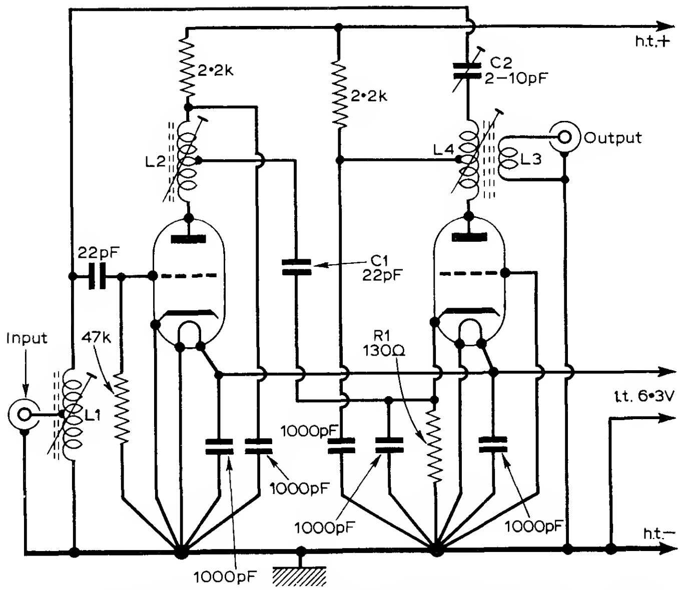


Fig. 5. A circuit using two Nuvistors in cascade, one as an earthed-cathode triode and the other in the earthed-grid mode.

This kind of circuit gives a slightly inferior mutual conductance due to the effect of the cathode bias, but is probably less critical from the stability point of view.

Note that inductive neutralisation may be used with leaky-grid bias and capacitive neutralisation with cathode bias. The inductive neutralisation and cathode-bias combination is shown in Fig. 4 simply for comparison with the opposite combination in Fig. 3.

Inductive neutralisation has several design problems, one being that the very low grid-to-anode capacitance of the valve demands rather a lot of neutralising inductance. Another is that it is often difficult to avoid coupling between the neutralising inductor and the tuning coils, especially in a small compact chassis with closely positioned components.

If there is coupling between the two circuits, the grid-to-anode feedback may be increased rather than decreased (neutralised), and great problems of instability will result.

From the experimenter's point of view, capacitive neutralisation is invariably the best bet. The neutralising capacitor should be of the air-dielectric variety, and the concentric type of trimmer is ideal for this purpose.

A Two-Nuvistor Circuit

Fig. 5 shows how two Nuvistors may be connected in cascade. The first is arranged in the earthed-cathode mode with leaky-grid biasing, as in

Fig. 2, while the second is wired in an earthed-grid circuit with the amplified signal tapped off L2 at a suitable point to match the cathode circuit of the second valve. Ordinary cathode-bias is used on the second stage, and the neutralisation is applied capacitively from the end of L4 to the grid circuit of the first stage, via C2.

Cascaded circuits are rarely required, however, for a single-stage Nuvistor circuit operating in front of a television tuner will almost certainly give a 2–3dB improvement in the noise figure while also providing a gain up to 40 or 50 times, depending upon the bandwidth and how well the amplifier is constructed and neutralised.

Chapter Sixteen

A CONVERTOR FOR TV SOUND

THIS small and completely self-contained unit enables the family radio receiver to perform an additional function—that of reproducing BBC-1 TV sound signals. Where a good quality broadcast receiver exists and *provided it is not of the a.c./d.c. variety* and it is reasonably sensitive, it is possible to make use of this for TV sound by changing the signals to a frequency covered by the receiver and injecting them into the aerial socket. The unit described here does this by converting the transmissions radiated in Band I to 1·5Mc/s. The signals are then fed to the aerial socket of a standard superhet and the resulting audio is heard at more than adequate volume via the loudspeaker of the radio set, quality being surprisingly good despite the necessarily narrow bandwidth of the i.f. amplifier. Although the signal/oscillator frequency is low, no operational difficulties should be experienced.

Some drift occurs when the unit is first switched on but may be catered for by suitably retuning the oscillator, for which purpose a reduction drive mechanism (to be described later) is fitted to the front panel.

The Circuit

This is given in Fig. 1 where a high-slope pentode amplifies the received signal. Use of this stage is desirable not only to give added gain but also to act as a buffer to prevent oscillator radiation that may otherwise reach the aerial. Where the gain is too great, as may occur in an area close to the transmitter, control may be effected by fitting a 5kΩ or 10kΩ wire-wound variable resistor in series with R4 at its chassis end. L2 is tuned partly by means of its core and partly by means of TC3, which is effectively connected across it via C3.

V2A operates as a conventional additive mixer, the requisite non-linearity being introduced by means of R6, etc., oscillations being introduced from V2B via C7. In some cases sufficient inter-electrode capacitance may make C7 unnecessary, in which case it may be omitted.

The oscillator circuit is so arranged that the cathode of V2B is tapped into the tuning inductance, since TC2, C11 and VC1 form a capacitive potentiometer across the coil L1. Both sections are adjustable, VC1 being connected to a panel control. The purpose of C11 is mainly to reduce the maximum capacitance value of VC1, but where a variable capacitor of 30pF is available this may be used instead of the two items shown in the diagram.

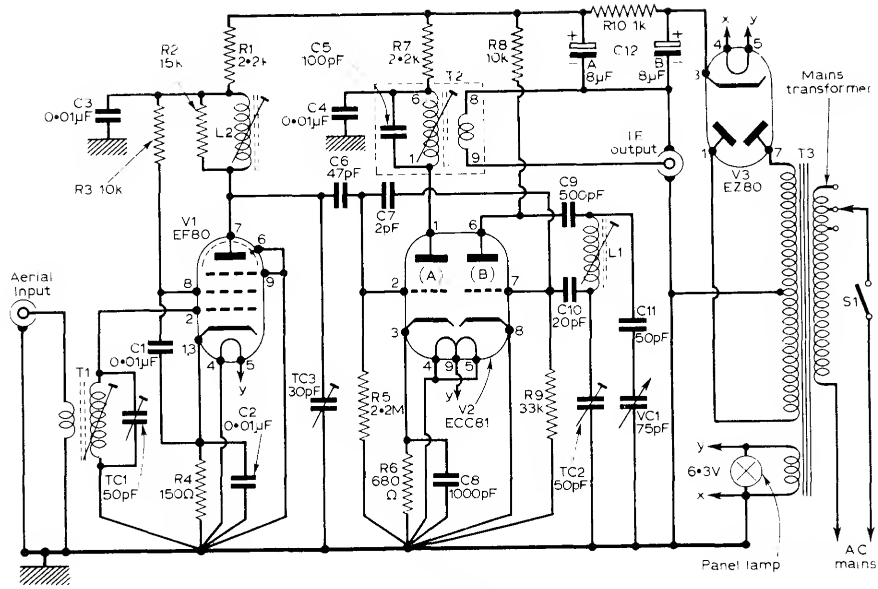


Fig. 1. The circuit of the unit.

Although, theoretically, C9 may be omitted, its inclusion is desirable to prevent d.c. from reaching TC2, which could, if faulty, short-circuit the h.t. supply to chassis.

Frequency

The oscillator may operate above or below the signal frequency and when the coils are suitably tuned an intermediate frequency of 1·5Mc/s appears at the anode of V2A and is developed across the main winding of transformer T2 which is tuned to this frequency. The impedance at this point is not suitable for connection to the receiver with which the unit is used, necessitating use of a low impedance winding. When the impedance transformation is complete the signals are conveyed to the output socket and thence via coaxial cable to the main receiver.

Two important points noted are: (1) T2 must be suitably screened, and (2) coaxial cable (television feeder) *must* be used to convey the signals to the receiver. If these points are not observed direct pick-up of unwanted signals at the converter i.f. (1·5Mc/s) is likely and will give rise to interference. Fortunately, screened coils are available and T2 consists of a Denco Yellow miniature coil, range 2 (medium-wave type). These coils are supplied in round metal containers designed to act as screening cans. They are also fitted with adjustable dust cores and a fixed capacitor, C5, is connected across the main winding instead of a trimmer, final tuning being accomplished by means of the core.

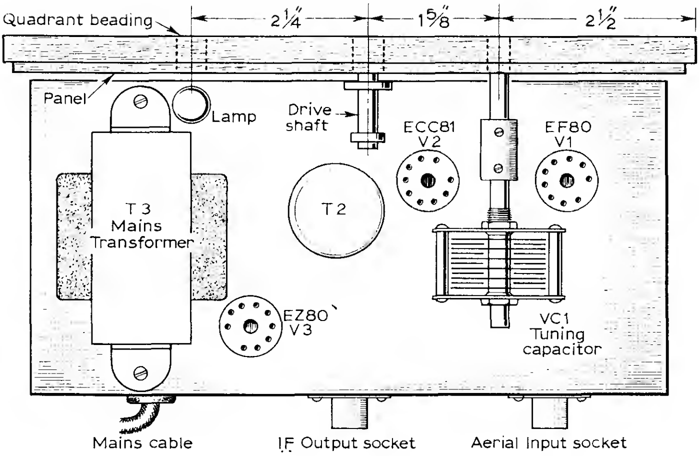


Fig. 2. The above-chassis layout.

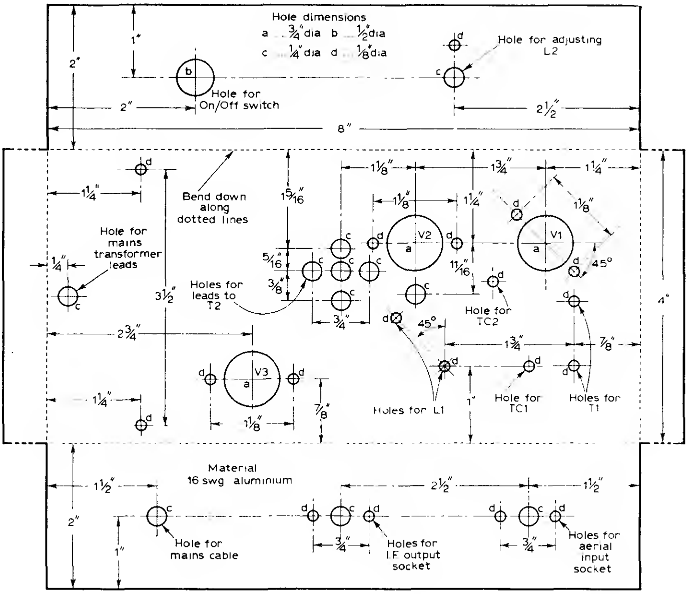


Fig. 3. The drilling details of the chassis.

Power for the Unit

Although, in some cases, supplies for the converter may be taken from the main receiver, it is generally best to use an independent supply provided by a separate or built-in unit. The latter method was adopted here partly for the sake of completeness and partly because a transformer with a secondary voltage rating of 180V-0-180V was available to provide full-wave rectification. A half-wave type pre-amplifier transformer can be used instead, but

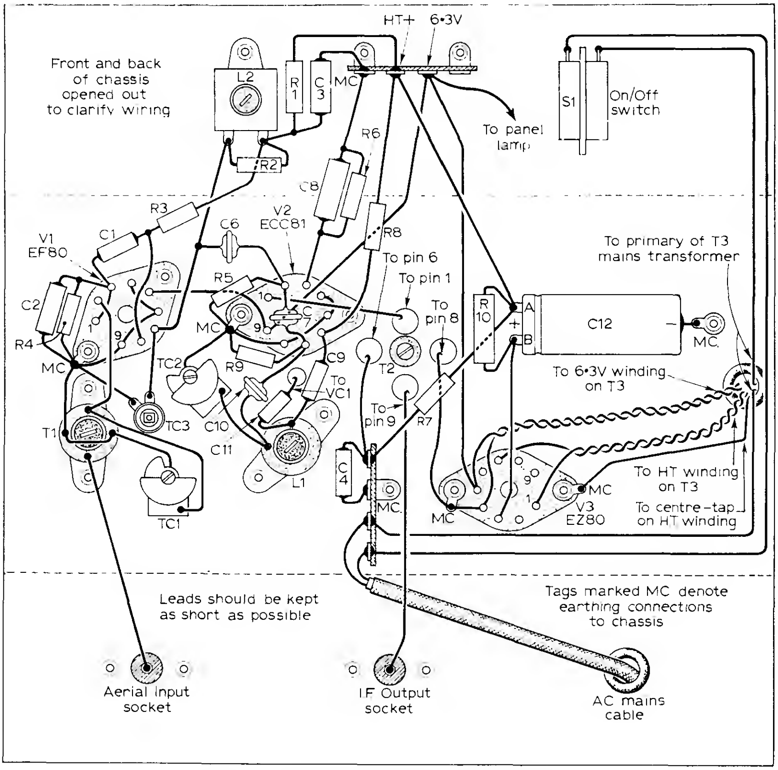


Fig. 4. The underchassis wiring diagram.

in this case, pins 1 and 7 of V3 should be wired together. An alternative is to dispense with V3 altogether, using in its place a small metal rectifier, but for various reasons the method adopted is preferred. The complete power section is shown on the right in Fig. 1, and the components are kept at one end of the chassis.

The chassis layout and drilling details are depicted in Figs. 2 and 3 and the power supply components occupy a distinct position so that anyone who does not want to include the section may quite easily omit it and use a chassis suitably smaller in size without in any way affecting the rest of the arrangement.

As may be seen from Fig. 4, nearly all the wiring has been kept below chassis and is so compact that no signal-carrying lead is longer than 1in; most are shorter than this, for it must be noted that in Fig. 4 the front and rear flanges of the chassis are shown pressed out flat in the interests of clarity. Note that short, direct wiring is essential in apparatus such as this designed for high frequencies.

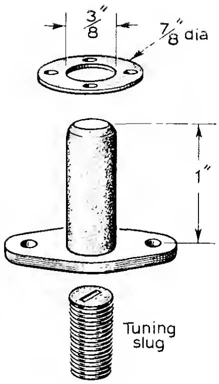


Fig. 5. The type of coil-former used in the unit.

TABLE A
Coil details: 30 s.w.g. enamelled copper wire
(Turns spaced by diameter of wire)

Channel Number	Number of turns			
	T1		L1	L2
	Sec.	Pri.		
1	6	$2\frac{1}{4}$	5	6
2	$5\frac{1}{2}$	$2\frac{1}{4}$	$4\frac{3}{4}$	6
3	5	$1\frac{1}{2}$	$4\frac{1}{2}$	$5\frac{1}{2}$
4	$4\frac{1}{2}$	$1\frac{1}{2}$	4	5
5	$4\frac{1}{2}$	$1\frac{1}{2}$	$3\frac{3}{4}$	5

The coils, which are wound on dust-cored formers of $\frac{3}{8}$in diameter fitted with push-on tag-rings, are positioned so that they may be removed and replaced easily without upsetting the rest of the wiring, so enabling adjustments to be made at the setting-up stage if found necessary. For details of the coils see Table A and Fig. 5. The trimmers are also easily accessible. T2 is retained by a 0B.A. polystyrene locking nut, the externally threaded section of the coil stem being passed first through the lid of the coil-container and then through the chassis so that the container body may eventually be turned upside down over the coil and screwed into the lid to form a rigid screening can.

Four small holes are also needed to carry the leads from the coil spills, the identification of which may be obtained from Fig. 1. A panel is bolted to the front flange and carries the warning lens, on/off switch and drive mechanism, a frame of $\frac{3}{8}$in quadrant being glued to the top, bottom and sides and allowed to overlap by $\frac{1}{8}$in to form a rabbet for the simple cabinet sections, dimensions of which are detailed in Fig. 6. Plywood or hardboard may be used for this, the inside of the panel being lined with metal foil connected to chassis.

Reduction Drive

This is very simple and effective. A standard tuning-drum of the type used in conjunction with a cord drive is attached to the end of the tuning capacitor spindle via a short length of $\frac{1}{4}$in rod and a coupler, these two items being necessary because the drum is mounted with its locking bolts close to the panel (see Fig. 7), which makes them inaccessible otherwise. The drum rim, grooved to accept cord, here conveniently works in conjunction with a

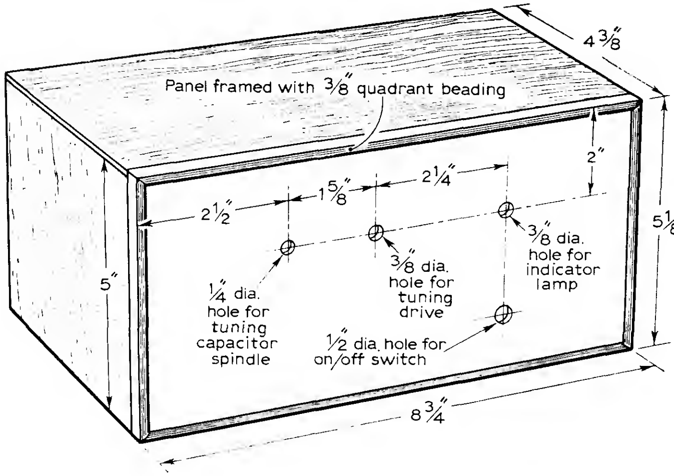


Fig. 6 (above). *Dimensions of the panel and cabinet.*

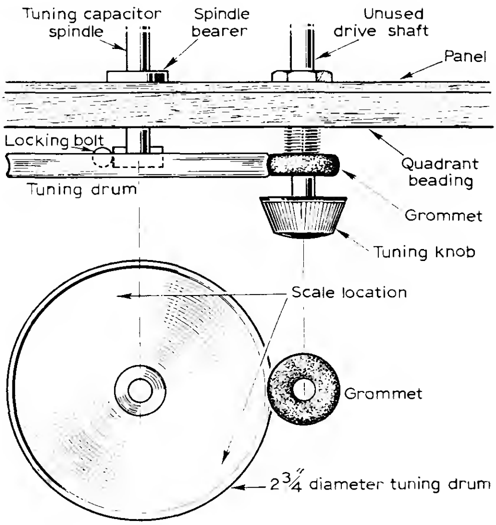


Fig. 7 (right). *Details of the tuning-mechanism.*

rubber grommet forced over the control spindle of a cord-drive shaft which is brought close enough to the drum to cause sufficient friction for it to be rotated. Provided the grommet is of a suitable size, fits tightly and is held securely by the shaft, a non-slip reduction drive of approximately 6:1 is secured, but the use of a ¼in diameter metal bearer for the tuning capacitor spindle is desirable to prevent it from becoming distorted.

Setting Up

On completion, a length of coaxial cable is fitted with aerial and earth plugs to the inner conductor and the braiding respectively at one end and a

standard coaxial plug at the other. The converter and the receiver are then interconnected and both switched on. The receiver is then adjusted to a quiet point on its dial around 200m and the volume control turned up, when a loud crackle should be heard via the loudspeaker when a penknife blade is scraped across the aerial-input socket of the converter.

Few constructors are likely to possess a signal generator operative at the frequencies to be received, so trial and error must be adopted. The TV aerial is plugged into the converter and an attempt made to obtain a signal by rotating the tuning capacitor, but, if silence obtains, as is most likely, VC1 should be set to mid-travel and TC2 adjusted very slowly and carefully, when it is possible that a weak signal will be obtained. This should be strengthened either by trimming the core of T2 or by moving the receiver pointer slightly, or both. If the signal sounds very distorted it is quite likely to be an f.m. transmission, thus indicating that the converter is tuned to too high a frequency. The core of L1 should then be screwed in a little more and a fresh attempt made.

A certain amount of patience is needed and it will be found a good plan to use TC2 to search for the vision signal (which will be heard as a low-pitched buzz) instead of the sound signal, with VC1 set so that its vanes are almost disengaged. If no signal is obtained when TC2 has been fully rotated, T1 and L2 should be adjusted slightly and another attempt made. Once the characteristic sound of the vision signal is heard, VC1 may be rotated slowly to a greater capacity, when the sound signal should be heard, and this may then be peaked up by using TC1, TC3 and the appropriate cores, not forgetting the core of T2.

If the converter fails to area operate in an where a signal may normally be received, switch off and insert a meter set to read 0–50mA between R8 and the h.t. line. Switch on and observe the current reading and if this is not excessive, switch the meter to read 0–10mA. Next, short-circuit L1 momentarily with a metal tool and note if the meter reading increases. If it does all is well, but if no change is detected switch off and investigate the oscillator and its circuit as this must be faulty. The fact that the oscillator might perform well at certain settings of TC2 and VC1 and not at others should be remembered when making the above test.

When the oscillator section is functioning correctly the meter may be disconnected and a fresh attempt made at receiving transmissions.

Correct Tuning

Incidentally, the operation of the unit can also be tested by switching the receiver to a quiet section of the long waveband or to around 600kc/s on the medium waveband and, once operation is obtained, little difficulty will be experienced in manipulating the transmission to the 1·5Mc/s region by moving the receiver pointer in step with TC2. On completion a scale should be drawn up on white card and glued to the tuning drum.

LIST OF COMPONENTS

Resistors:

R1	2·2kΩ ½W
R2	15kΩ ½W
R3	10kΩ ½W
R4	150Ω ½W
R5	2·2M ½W
R6	680Ω ½W
R7	2·2kΩ ½W
R8	10kΩ ½W
R9	33kΩ ½W
R10	1kΩ 3W

Capacitors:

C1	0·01μF ceramic
C2	0·01μF ceramic
C3	0·01μF ceramic
C4	0·01μF ceramic
C5	100pF ceramic or mica
C6	47pF ceramic or mica
C7	2pF ceramic
C8	1000pF ceramic or paper
C9	500pF ceramic or mica
C10	20pF ceramic or mica
C11	50pF ceramic or mica
C12	8+8μF 275V elec.
VC1	75pF variable (*see text*)
TC1	50pF ceramic trimmer
TC2	50pF ceramic trimmer
TC3	30pF concentric trimmer

Valves:

V1	EF80
V2	ECC81
V3	EZ80

Valveholders:
Three B9A

Mains Transformer (see text):
Mains A.C. input
Secondaries—180V–0–180V 35mA; 6·3V 2A (a transformer with an h.t. secondary of 200V–0–200V or 250V–0–250V may be used if R10 is increased so that the h.t. produced is about 180V–190V).

Coil Formers with Tag Rings and VHF-grade Dust Cores:
Neosid or Aladdin, ⅜in diameter

T2:
Denco Yellow medium wave coil, Range 2

Chassis:
16 s.w.g. aluminium, 8in × 4in × 2in

Tuning Drum:
2¾in diameter and Cord Drive (see text)

Switch:
Mains on/off type

Miscellaneous:
Spire clip; tag strips (2); grommets; wire; spindle coupler; nuts; bolts; tags, panel and cabinet material; coaxial cable; two coaxial sockets, etc.

Chapter Seventeen

TELEVISION TEST CARDS "D", "E" AND "F"

ALTHOUGH Test Cards D and E are similar to Test Card C, a number of changes take into account improvements in television equipment in recent years. Test Card D is used on 405-line transmissions, and Test Card E on 625-line transmissions. Each pattern on the cards is designed to assess particular characteristics as follows:

Aspect Ratio

The central concentric black and white circles should appear truly circular when the width and height of the picture are adjusted to the standard aspect ratio 4:3.

Adjustment of Picture Size

The limits of the transmitted picture are indicated by the point of contact of opposing arrow heads on each side of the test card and by the outer edge of white squares in each corner. As most receivers have a display area aspect ratio of approximately 5:4, it is usual to adjust the receiver so that the top and bottom edges of the display area coincide with the arrow heads of the test card and the side castellations of the test card just appear in the display area of the receiver.

Resolution and Bandwidth

Within the circles is a group of six frequency-gratings each consisting of alternate black and white vertical stripes corresponding to fundamental frequencies (in Mc/s) of:

Test Card D		Test Card E	
1·0	1·5	1·5	2·5
2·0	2·5	3·5	4·0
2·75	3·0	4·5	5·25

The gratings are designed to produce after gamma correction a signal of approximately sine waveform corresponding to 50% modulation (as opposed

to the square waveform on the earlier cards). Thus the stripes have a gradual transition from black to white.

A new version of Test Card D came into use late in 1965 (Test Card D is used on 405-line transmissions). The new card is for the most part the same as the original version but the range of brightness in the frequency gratings has been increased. The modified card is identified by a white dot on each side of the letter D near the bottom.

The six frequency gratings consist of vertical stripes designed to produce, after gamma correction, signals of approximately sine waveform, corresponding to the following frequencies (as before): 1·0Mc/s, 1·5Mc/s, 2·0Mc/s, 2·5Mc/s, 2·75Mc/s and 3·0Mc/s. The range of brightness in the gratings is now the same as that from the first (top) square of the contrast pattern to the fourth square. The brightest parts of the stripes have the same brightness as that of the area surrounding them.

Contrast

A five-step contrast wedge appears in the centre of the test card corresponding to a contrast range of about 30 to 1 between the black and white squares at each end of the scale. Adjacent squares should give equal brightness difference on a correctly adjusted receiver. Within the top and bottom squares are small circular areas of slightly brighter tone. The merging of these areas into the surrounding area indicates white or black crushing as the case may be.

Scanning Linearity

The background of the test card is a medium grey, bearing a graticule of white lines. These should be reproduced in all parts as enclosing equal squares and the central black and white rings should appear truly circular.

Line Synchronisation

The alternate black and white rectangles forming the border serve to check the line synchronisation of receivers. Faulty line sync will show as horizontal displacement of those parts of the picture following the white rectangles, in particular giving the central circles an appearance of "cog-wheels".

Low-frequency Response

Poor low-frequency response will show as streaking at the right-hand edge of the black and white areas of the rectangles at the top centre of the test card, and also at the right of the border castellations.

Reflections

The white vertical line with the black background and the black vertical line with the white background should appear free from images (ghosts). These lines represent pulses having a duration of 0·3μs on Test Card D and 0·2μs on Test Card E. Reflections from hills or large buildings may result in displaced "ghost" images. This effect will be most readily seen as displaced positive or negative images of the vertical lines mentioned above.

Uniformity of Focus

In each corner of the test cards there is a diagonally-disposed area of black and white stripes; the focus of these areas and the central area of the test card should be uniform. The stripes correspond to fundamental frequencies of about 1Mc/s on Test Card D and about 1·5Mc/s on Test Card E.

Test Card F

During 1967, a new test card came into use on 625-line transmissions. The new card, known as Test Card F, was designed jointly by the BBC, B.R.E.M.A., E.E.A. and I.T.A., and provides tests for the various functions of colour and black-and-white television receivers.

Other features of Test Card F fulfil similar purposes to those on Test Cards D and E. However, the frequency gratings are equivalent to square-wave signals of the following frequencies (from top to bottom): 1·5Mc/s; 2·5Mc/s; 3·5Mc/s; 4·0Mc/s; 4·5Mc/s; and 5·25Mc/s.

Section 5

TEST GEAR

Chapter Eighteen

THE VIDEOSCOPE OSCILLOSCOPE

THE Videoscope is built around the surplus VCR139A c.r.t. which is obtainable quite cheaply. Whilst these tubes are often of somewhat poorer quality than commercial types, the cheap price more than compensates for this. Considerable experimenting to obtain a circuit highly tolerant of c.r.t. shortcomings led to a satisfactory prototype.

At the low e.h.t. voltages cheaply obtainable from slightly modified conventional h.t. supplies, the VCR139A suffers from astigmatism and deflection-defocusing to an extent greatly exceeding the tolerances of even the most broad-minded reader, unless push-pull deflection is used throughout, on both X and Y plates.

Some tube specimens were found better than others, but even then push-pull deflection is still required to obtain the necessary peak-to-peak deflection amplitude of nearly 200V. Single-valve deflection would require rather a large h.t. voltage to generate this output, with consequent problems and expense in components—particularly electrolytics.

To avoid the expense, weight and space of numerous iron-cored smoothing chokes, yet maintain excellent h.t.-feed decoupling of the various portions of the instrument, smoothing resistors of rather large values compared to the currents drawn are used, thus necessitating a relatively large h.t. input from the rectifier (350V).

It is vitally important *not* to substitute a metal rectifier for the EZ80 valve-rectifier, V8, as the heater warm-up time of the rectifier gives the essential delay in application of voltage to the electrolytics until all valves have warmed up.

With a metal rectifier there would be an immediate surge to about 500V on the h.t. line when switching-on, which would remain until the valves had warmed-up, a time sufficient to destroy the electrolytics.

Constituent Parts of Circuit

The Videoscope is composed of four parts. The first is a high-gain Y-amplifier (Fig. 1), with capacitive and inductive compensation in a form which can easily be managed by the normal constructor. This gives full amplification not only over all audio frequencies of the hi-fi scale, but

also through the entire long and medium wavebands of radio frequencies and some way beyond.

The amplifier is, of course, not tuned to any of these frequencies—it passes and amplifies all of them linearly; it is, in other words, a true video amplifier with the necessary push-pull output for symmetrical deflection of the c.r.t.

The effective sensitivity at the input terminal is so high that direct pick-up of local medium-wave broadcast stations, without any leads or aerial connected whatsoever, can lead to full screen-size display of the modulated carrier(s) on the c.r.t. in areas of high signal strength, making the most sensitive setting completely unusable unless efficient screening is used.

The sub-chassis space is partitioned into compartments for the Y-amplifier and the remaining circuitry. This is to avoid radiation of r.f. from the timebase circuitry into the Y-amplifier input (flyback transients of the timebase waveform, especially at the higher speeds provided, have an r.f. spectrum).

It is emphasised that the compartment-screen is *essential*; if omitted, the Y-amplifier is unstable and permanently blocked by injected timebase transients. This instability and interference persists even with the compartment screen, until an aluminium sub-chassis cover, details of which will be given, is fitted. The circuit is then quite stable, with no movement of the trace in the most sensitive position when no Y-input is connected.

The sub-chassis cover is effective in preventing entry of medium-wave broadcast signals. In this respect it is also essential to use skirted valveholders fitted with screening cans entirely enclosing the valves. This further prevents broadcast pick-up and interference radiation between valves above chassis. Only the rectifier valve should be unscreened (for cooling).

These measures are together most important, so that only those signals deliberately fed to the coaxial Y-input socket are displayed. The deliberate input may be audio, i.f. or r.f. as desired.

In the latter cases, no detection takes place, of course, i.e. the carrier-train is shown as such on the c.r.t. screen. If modulated, the modulation envelope is visible, and the percentage modulation can be determined visually.

For this purpose, the timebase must run at a low audio-frequency (between 10c/s and 25c/s) for general visibility of mixed music and speech modulation. For a fixed-tone modulation-note, the timebase may be synchronised to a sub-harmonic in the usual way; the sync-amplifier V4, thereby automatically functions as detector (power anode-bend type) to generate a sync-signal corresponding only to the audio modulation. If the probe (described later) is connected across a simple tuned circuit (externally), various experiments on aerials, tuned circuits, modulation-adjustment of signal generators, etc., are possible. This is just mentioned as an addition to all the "normal" uses of an oscilloscope for displaying waveforms, which need no further comment here.

The Probe and Attenuator for the Y-amplifier

It is highly useful if the amplifier can be used for quantitative voltage

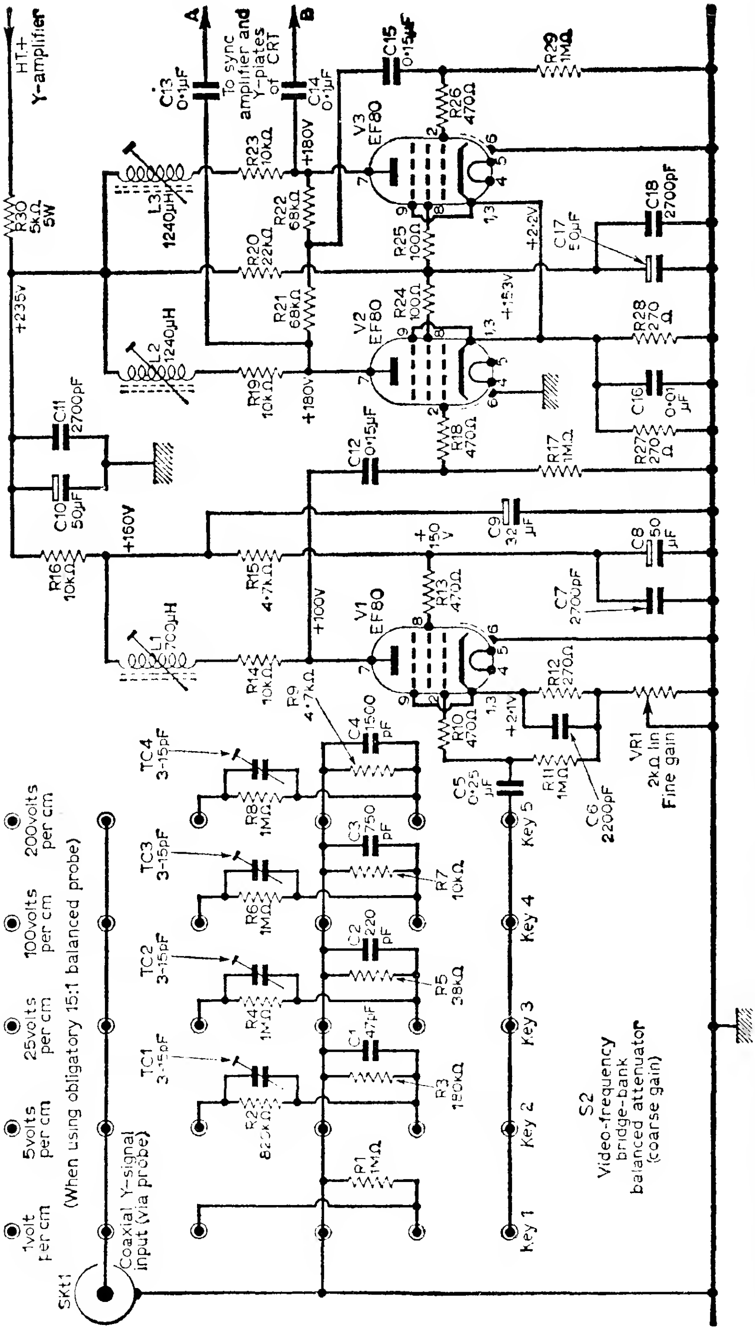


Fig. 1. The circuit of the Y-amplifier, which has a response which is level from 5c/s to 1Mc/s. The amplifier operates to well over 2Mc/s and the gain is approximately 3000. The piano-key switch S2 forms the bridge-bank attenuator—each key constitutes a d.p.d.t. switch. In the "off" position, the centre contact of each set is made to the upper contact; in the "on" position (key depressed), the centre contact is connected to the lower contact. Only one key at a time may be depressed due to the mutual-release mechanism. The voltage measurements shown were made with a 4kΩ/V meter with no signal input and VR1 at maximum (slider at V1-cathode end).

measurements from audio to r.f. frequencies, over as large a range of input amplitudes as possible, ranging from a fraction of a volt to hundreds of volts (more than full-screen voltage required by c.r.t.).

A method of gain control is thus required which operates unambiguously from the highest possible initial maximum sensitivity down to a gain less than unity. The last condition is essential, as many waveforms arising in TV experiments have amplitudes of hundreds of volts, yet must be viewed, and thus accepted by the amplifier input. Furthermore, the frequency and phase characteristic of the amplifier must be level over the entire video range *at all gain settings.*

It is clear that a simple volume-control potentiometer is useless here; it cannot be set with reproducible accuracy over such a wide range, and its stray capacities cause variation of frequency response and phase response with change of setting.

The method of gain-control thus adopted consists of a bridge-bank balanced attenuator built around a five pushbutton aggregate (a row of five d.p.d.t. toggle switches is usable as a substitute), to select five basic sensitivities. A negative-feedback cathode-potentiometer in the input stage gives continuous fine-control in about 6 : 1 ratio, to cover the gaps between the five basic steps (see Fig. 1).

The only drawback of this combined arrangement, otherwise excellent, is that operation of the cathode potentiometer VR1 necessarily generates a spurious transient signal whilst turning, so that operation must be gentle to avoid brief flutter of the trace off the screen. This was considered tolerable after practical operation with the prototype; further balancing arrangements to overcome it were considered an unnecessary complication.

Signal Cable

It is necessary to feed the signal to be viewed from its point of origin to the Y-input socket of the Videoscope, and a yard or two of coaxial cable is required for this. The total stray capacity of this cable together with the input circuitry will lie around 200pF, i.e. it is too large to impose across the circuitry being tested.

It is thus essential to use a correction-probe head at the front end of the coaxial cable. This consists simply of a parallel combination of high value resistors and small ceramic capacitors, forming a balanced-bridge bleeder, raising the input resistance and lowering the input capacity. The proportionate loss of overall gain is the price paid for this, but is tolerable on account of the very high initial gain of the amplifier.

It is of great advantage to make the V/cm calibrations on the Y-gain controls correct for, and including, the characteristics of the probe. One will only in very exceptional circumstances operate without the probe, so that its inclusion in the calibration avoids constant mental arithmetic when working with the unit.

Inclusive of the probe in this way, the Videoscope is calibrated for a

continuous coverage from 333mV/cm to 400V/cm at the Y-amplifier input. Without the probe, i.e. for direct input at the coaxial socket, the sensitivity is 15 times greater but the frequency and phase correction is lost above audio frequencies in general.

Together with the probe to be described, a waveform of maximum peak-to-peak amplitude of 1·2kV can be accepted and displayed, and a d.c. voltage of 1·5kV can be blocked on all ranges. This enables measurements to be made with safety on waveforms tied up with the boost-voltage feeds of TV receivers.

D.C. voltage measurements are made by repeatedly jabbing the prod of the probe against the point to be tested, with the timebase running at the slowest speed provided (20msec/cm). Owing to the excellent bass response of the Y-amplifier, a series of square pulses will then be observed, their height giving the d.c. voltage value (same calibrations as for a.c.) and their direction the polarity.

D.C. voltages as low as 10mV can be measured with reasonable accuracy (direct input without probe), and up to 1·5kV (with probe) at the other extreme, with source-loading characteristics similar to valve-voltmeters, i.e. much less than with multimeters.

The Timebase Circuits

The timebase is a stable and reliable circuit (Fig. 2). Experiments were made using other types of timebase arrangement, particularly the two-valve Sanatron. This can be triggered instead of synchronised, by applying suitable cut-off bias, and for such experiments R71 (Fig. 8) was originally placed below chassis, to make negative voltages available there. This is not required in the final design, but is useful if the advanced experimenter wishes to try other timebases. Grommet 5 may be omitted and R71 wired across C52 above chassis. It was left in the sub-chassis position in the prototype after stripping out the original Sanatron arrangement and replacing with the Miller-transitron design. The Sanatron has the disadvantage that it needs a pair of fine-speed controls (a rundown and a stability-hold), which must be critically adjusted to remove the bright waiting-spot at the end of the trace, yet not silence the oscillator. The adjustments of these two controls in the simple design tried were coupled with much backlash, so that performance was not sufficiently elegant. Furthermore, trigger and sync functions over the entire speed range were difficult to arrange with uniform stability without introducing complications involving further valves.

The only advantage of the Sanatron over the Miller-transitron is its ability for trigger operation, i.e. non-free running, giving one cycle only per Y-signal or other input cycle. This allows a very fast speed to be set, even when the repetition rate (which may be irregular) is very low, so that, as a result, transient oscillations and effects are greatly magnified in time-scale.

The usefulness of a trigger operation is greater the faster the transient run and the longer the waiting time between, as this ratio gives the "time

magnification" achieved compared to normal sync operation. But in the same ratio, the screen brightness of the display on the c.r.t. is reduced, as the stop is absent for an increasing percentage of the time.

The trigger-displays with the VCR139A were, accordingly, very faint, and needed a darkroom for observation. Thus, apart from the shortcomings of a wide-range Sanatron regarding stability in simple arrangements, it was realised that the VCR139A is basically unsuitable for this mode of operation at high transient ratios—a successful trigger-oscilloscope requires a tube of high intrinsic brilliance, preferably with post-deflection acceleration of several thousand volts.

The intrinsic brilliance of most samples of VCR139A is satisfactory for normal sync operation with a continuously running timebase, but nevertheless even there rather low. The e.h.t. voltage used in the present design may be taken as the absolute minimum usable whilst maintaining acceptable brilliance. It was chosen to cut expenses of h.t. and e.h.t. supplies.

Timebase Speeds

Pressing for simplification wherever possible in the present design, unusually large coarse-speed control steps (more than 10:1) have been adopted, reducing the number of expensive accurately matched pairs of timebase capacitors needed.

This is an unusual procedure in oscilloscopes, because it means that, if high total speed range is thereby maintained, the flyback and forward times are about equal at the fast end of each range on the fine-speed control. This is of itself no disadvantage, but the consequent problems of accurate flyback blanking at the fastest speeds and fast ends of the ranges are thereby aggravated.

The screen-grid waveform of a Miller-transitron is an approximate square-wave, positive half-waves thereof roughly coinciding with the forward stroke and negative half-waves with the flyback of the anode waveform. This waveform is applied, in principle, to the c.r.t. grid, with its positive level clamped with a diode to the slider of the brilliance-control. This ensures that the set-brilliance voltage is held during all strokes, and the c.r.t. is cut-off at the grid during flyback. Now it is a fact that the negative-going stroke of the screen wave, at the end of the timebase forward stroke, is generally sharp, giving immediate extinction of the c.r.t. trace at the very start of the flyback, even at high speeds. But the positive-going stroke to bright-up the c.r.t. trace at the start of a new timebase-run is generally much slower, taking an appreciable time to reach its final level. At high speeds this can lead to late bright-up, causing the first portion of the trace to remain dark, and thus be lost.

Apart from consideration of stray capacities, the precise characteristics of the pentode chosen for V5 play an important role, and in general valves with a good screen-to-anode gain are best.

It is also of advantage to choose a valve of low mutual conductance at the

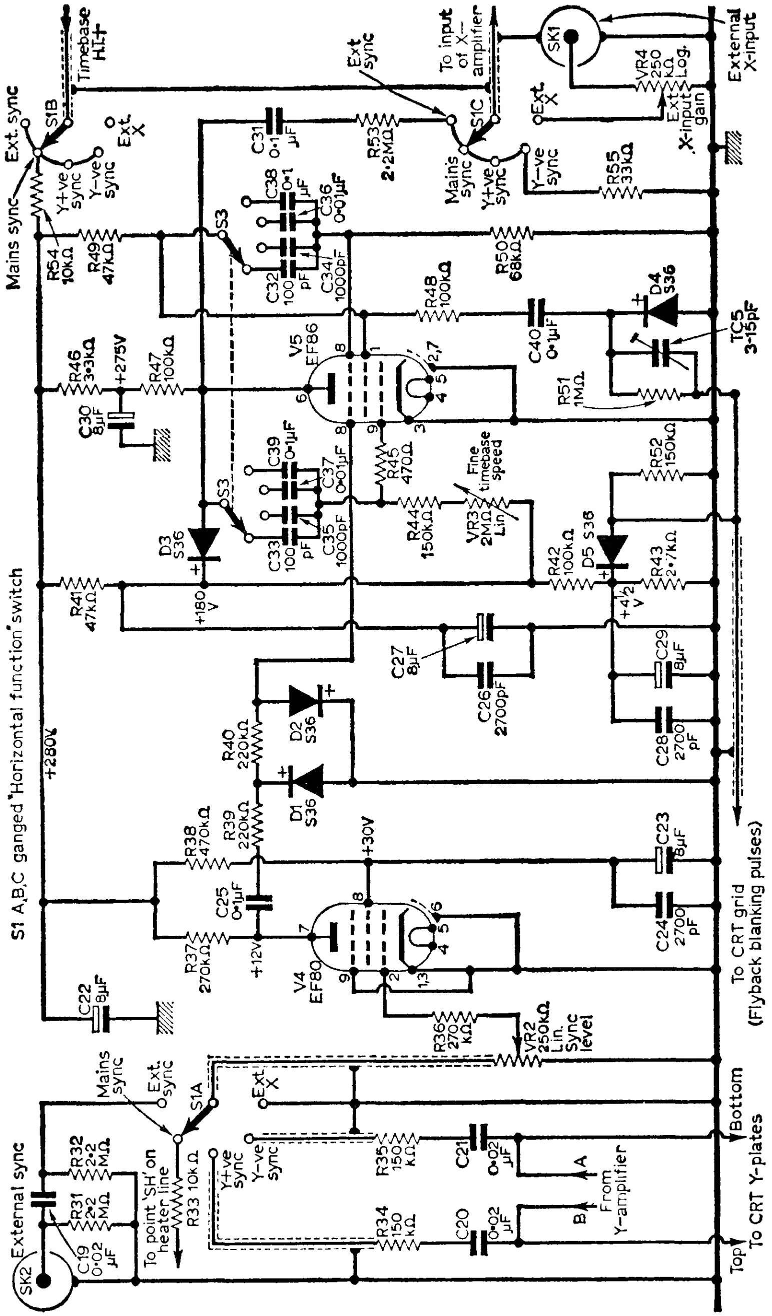


Fig. 2. Timebase and sync circuits. Four speed ranges and a fine-speed control are used (covering speeds from 2μs/cm to 20ms/cm) and a diode shaping-circuit obtains flyback-blanking pulses for the grid of the c.r.t. from the screen-grid of V5.

control grid, as this leads to larger values of required timebase capacitors at the higher frequencies (speeds), swamping the undesirable effects of stray capacities. This is the ultimate reason why an EF86 has been selected for V5, whereas all other positions are occupied by EF80's giving higher signal-gains at low anode potentials.

Flyback-Blanking

R48 (Fig. 2) is a series resistor to reduce screen loading of V5 by the blanking circuit. C40 blocks the screen-voltage d.c. component, and together with D4 clamps the waveform entirely positive-going with respect to chassis. R51 and R52 constitute a voltage-divider, to reduce the high amplitude to the lower amplitude required at the c.r.t. grid.

TC5 is the bridge-balancing capacity to offset the effects of the considerable capacity of the screened cable to the c.r.t. grid. A low-capacity type should be used for this cable for best results. D5 is a limiter diode, selecting only the lower, 4·5V of the attenuated waveform, thus rejecting the more rounded top, to offset the bright-up delays already mentioned. Finally, D8 (Fig. 3) clamps the 4·5V amplitude square-wave entirely negative-going with respect to the brilliance control slider.

TC5 can be adjusted empirically for optimum of brilliance uniformity at high speeds on the finished oscilloscope, or, more exactly, with a second oscilloscope observing the negative square pulses at the junction of R51 and R52. TC5 is first screwed to maximum capacity, and then reduced until the sharp spike (negative) at the start of each pulse just vanishes, leaving a clean square half-cycle.

Provided a good low-capacity cable has been used for the c.r.t. grid lead, unnecessary lengths and stray capacities in the associated circuitry right from V5 screen grid avoided, and TC5 has been properly adjusted, blanking precision is virtually perfect on three ranges, and only slight loss at the start of the trace is apparent on the fastest range over the final fastest 20% of its coverage.

If the screened cable is of really low capacity, the performance of the fastest range at the fastest end can be further improved by judicious increase of R52.

Timebase Speed Ranges

The timebase speed controls are calibrated in time/cm values, not in frequencies. This is far more useful and unambiguous than the antiquated form of frequency-calibration, because it allows quantitative time and frequency readings on viewed signals without falsification or ambiguity due to the unknown duration of the flyback.

If a signal waveform is locked with the sync control, and the length of one cycle noted against the centimetre-grid scribed on the Perspex window in

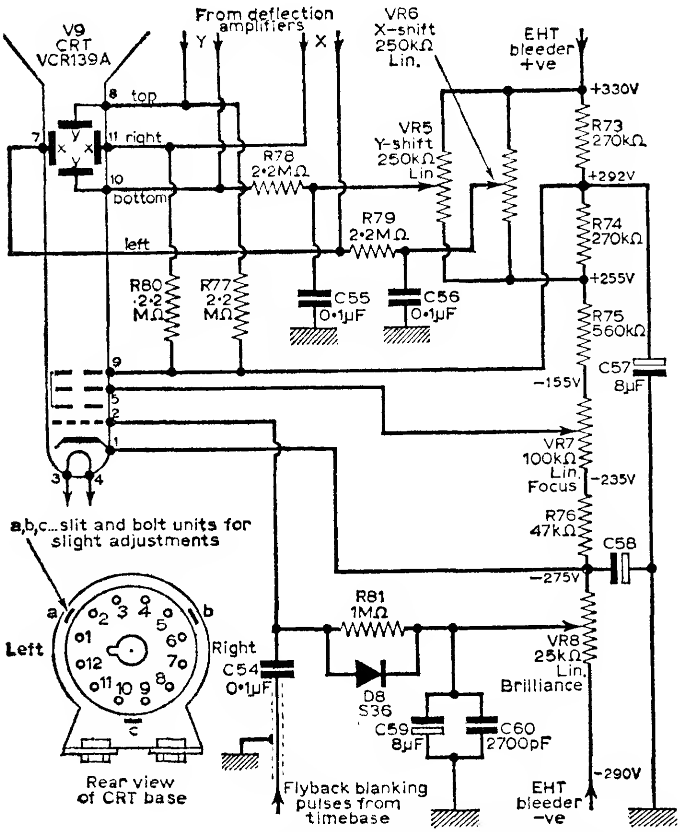


Fig. 3. The c.r.t. circuit. If the base of the c.r.t. (viewed from the rear) is orientated as shown and the connections given here followed, the trace will be of correct orientation and polarity. Note that the X and Y plates are interchanged for reasons explained in the text.

front of the c.r.t., then this length multiplied by the time per centimetre read off the speed-control's calibration gives the time for one period of the waveform being viewed, whose reciprocal is the frequency thereof, in an unambiguous fashion.

The fine-speed control is calibrated in time/cm units from 2 to 20; the coarse control (S3) gives four decimal steps, so that the "units" are milliseconds, hundreds of microseconds, tens of microseconds and microseconds respectively. This corresponds to a continuous coverage from about 5c/s to about 50kc/s timebase frequencies in the older nomenclature.

Referring to Fig. 2, V4 is a sync-amplifier with very high gain, so that a rigid lock of even small-amplitude signals is possible. S1 enables internal

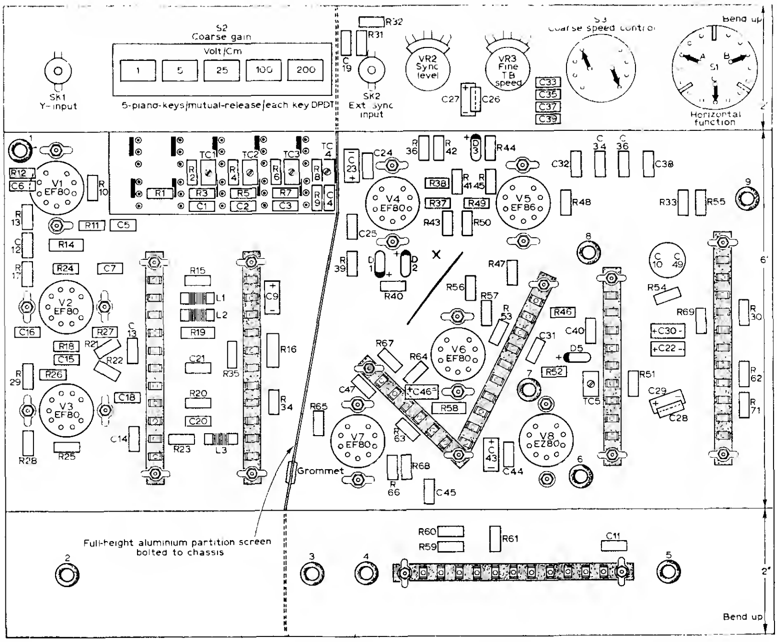


Fig. 4. The underchassis layout. The screen is essential to prevent blockage of the sensitive Y-amplifier by injection of transients from the timebase.

sync action to be selected from either the positive or negative transitions of the Y-signal or from the mains, or from an external sync signal fed in at SK2. In all cases VR2 controls the sync intensity. The setting of VR2 should be as low as possible, consistent with good lock, to prevent the generation of excessive transients at V4 anode which can radiate into the X-amplifier and cause peculiar kinks in the displayed waveform.

Excessive application of sync, greatly in excess of the amount at which rigid lock sets in, i.e. with VR2 turned up too high, generally leads to notches on sharp transients of the viewed waveform, which wander up and down these flanks in the direction of the arrows (Fig. 10) as the sync control VR2 is adjusted. If such injection-kinks do not vanish fully some considerable distance of travel of VR2 before lock is lost, then the screening at V6 grid is not properly adjusted. (Refer to Fig. 8 and especially to Fig. 4, wiring diagram.)

The thick line labelled "X" shown in Fig. 4 between V4 and V6 is an insulated "floating screen". This consists of a piece of thin brass foil cut somewhat larger than a postage stamp, with an insulated lead soldered to each edge. The whole is then wrapped in insulating tape, and the appearance is that of a small flat capacitor. The two leads are soldered to the chassis-solder tags shown. This insulated floating screen is then bent into the optimum

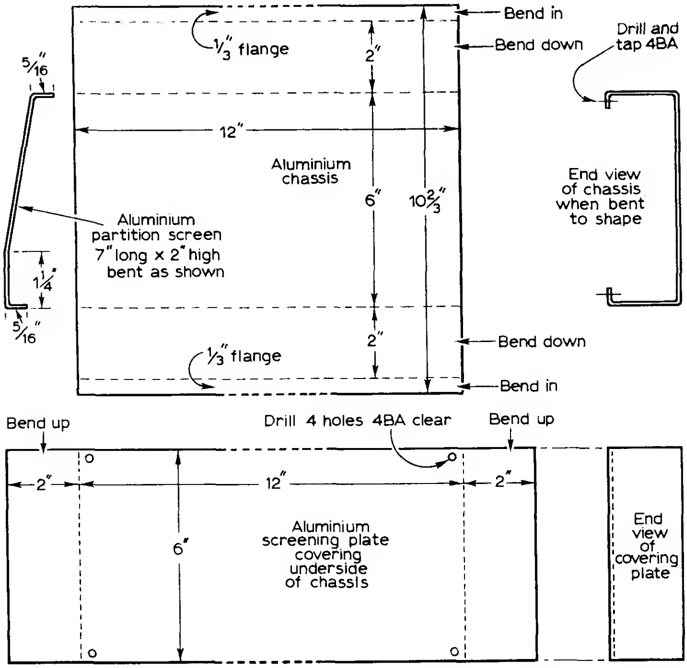


Fig. 5. The covering screens of the sub-chassis. These prevent pick-up of transients from the timebase and direct interference from pick-up of broadcast stations.

position in the vicinity of V6 grid (R56, R57) until all sync-transient radiation from V4 is kept out of V6 for normal settings of the sync control and a safe margin beyond. Such an optimum adjustment is possible because of the relatively low impedance of V6 grid circuit in all timebase-operation positions, because R55 is then in circuit. In the position for "External X-deflection" applied to SK3 and controlled by VR4, the grid-impedance at V6 is much higher—but the sync amplifier is muted then because h.t. is switched-off for it and the whole timebase circuitry, at S1b, thus removing possibilities of sync-transient injection at V6 grid.

Fig. 5 shows constructional details of the aluminium sub-chassis covering screens—not to be confused with the floating screen referred to above. The above-chassis layout is shown in Fig. 6, and the c.r.t. mounting and screen window details are given in Fig. 7.

The X-Amplifier (Fig. 8)

R55 together with R53, forms a voltage divider for the timebase signal from V5, giving the X-amplifier (V6 and V7) the correct drive. This form of voltage division is preferable to a simple tapping of R47 near the h.t.-end

for two reasons. Firstly, tapping R47 would certainly reduce the timebase voltage as desired, but would not give the slightest reduction of fluctuations on the h.t. line due to mains transients coming from outside or due to flickerings quite normal for all electrolytic capacitors and amounting to appreciable fractions of a volt. Such flickerings, which are random, receive 100-fold amplification in the X-amplifier, and would thus lead to continuous jitter of all displays by amounts up to 30% of full deflection, which would be intolerable. The arrangement of R53 and R55, however, reduces signal and flickering by the same amount, so that the so-called "flicker-ratio" is not increased. This gives a more stable display, essential if waveforms are to be photographed off the screen. The second advantage of the arrangement used is that high voltages are kept off S1c and the impedance there is low, so that stray capacities do not impair the action at high speeds.

Some attention must be given to trimming-up for optimum linearity in the X-amplifier. The critical stage is V6, as this is supplying all the gain.

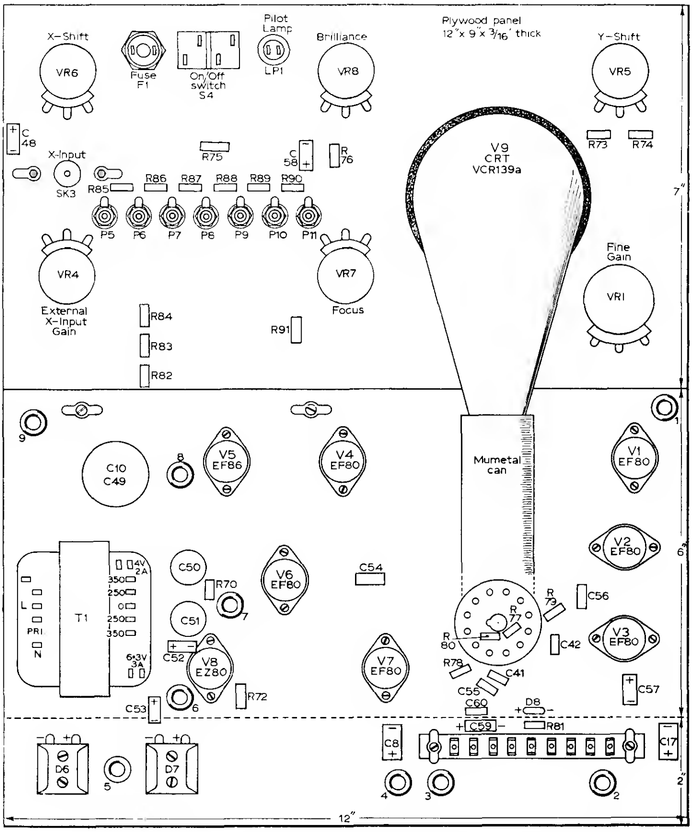


Fig. 6. Layout of the major components above the chassis.

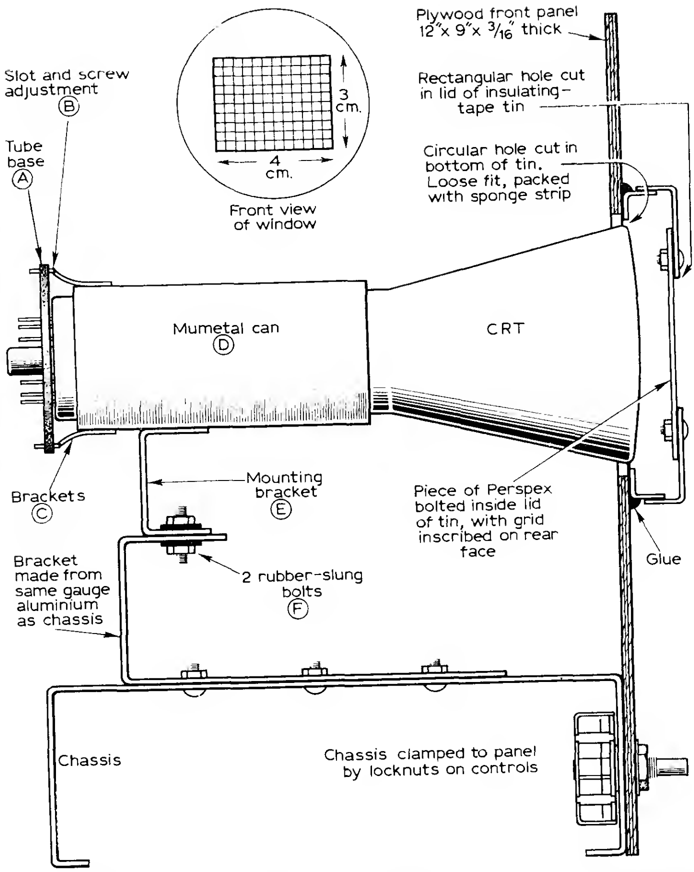


Fig. 7. Details of the screen window and of the mounting arrangements of the c.r.t.

V7 is self-linearising, being "slaved" at unity gain as phase-inverter. Similarly, the timebase valve itself is also self-linearising, by its very nature, so that when used direct in single-ended deflection, linearisation is automatic.

As V6 must operate linearly at an anode output signal of around 100V peak-to-peak, the adjustment of R67 is critical. A resistor of approximately the value specified should be selected and observation made on the c.r.t., at the same time feeding a sine-wave signal of about 1kc/s into the Y-amplifier and locking this with the sync. Lack of X-linearity is evidenced by uneven spacing of the displayed cycles. If satisfactory linearisation is not possible without undue departure from the specified value of R67, then negative feed-back must be introduced by inserting equal individual cathode resistors (unbypassed) between the top of R67 and the respective cathodes. Values between 10Ω and 27Ω can be tried, reducing R67 by *half* the value finally

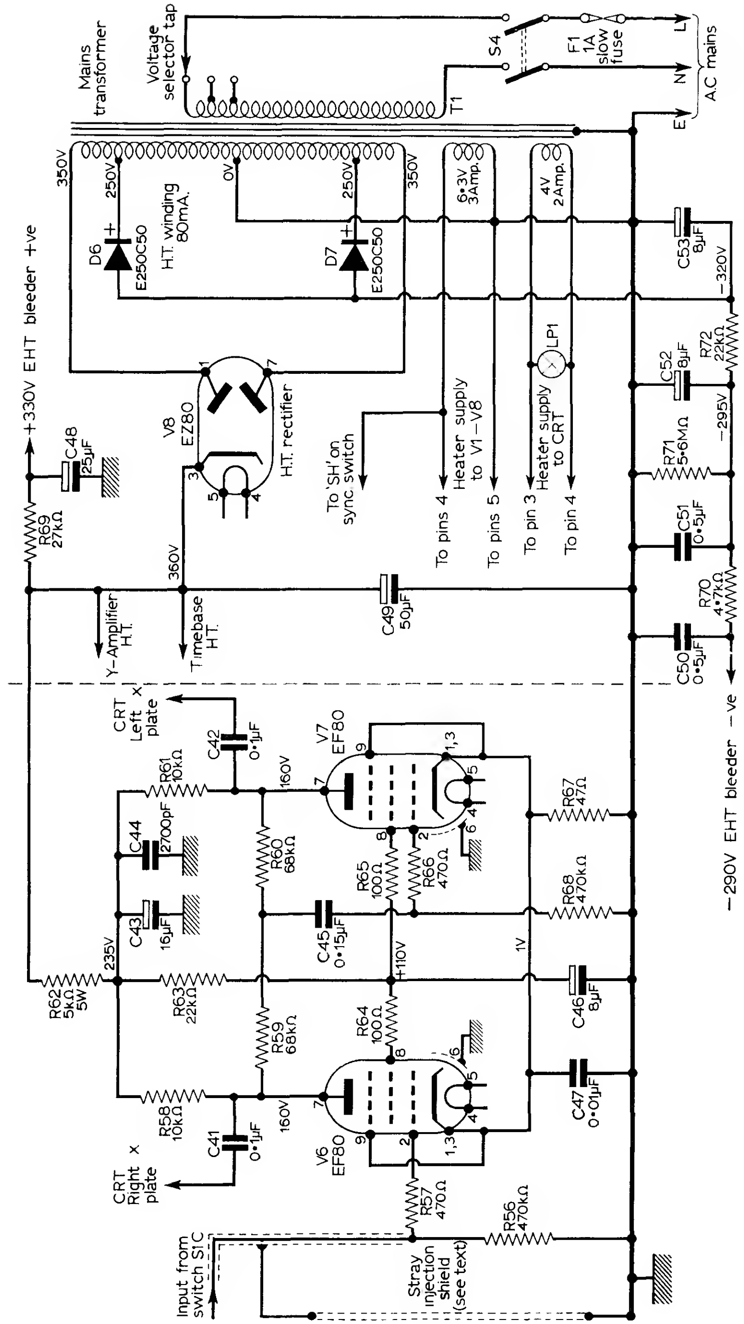


Fig. 8. The X-amplifier and the power supplies. Conventional components are used in the power supply. The X-amplifier can be used with an external signal, or with the built-in timebase of Fig. 2. To avoid h.t. surges, V8 may not *be replaced with a metal rectifier or silicon-diode system.*

selected here. The drive applied to V6 may be adjusted if necessary by increasing the value of R53. Do *not* reduce the value of R55 for this purpose, because this can lead to different values of grid-current bias for V6, V7, causing unequal anode currents.

If the anode voltages of V6, V7 (the X-amplifier, Fig. 8) depart from the values indicated in Fig. 8 or if more than 20V difference exists between the two anodes, adjustments should be made by suitable inequalities of the separate individual additional cathode-resistors already mentioned. R67 and any auxiliary cathode resistors must not be reduced below values giving 10mA standing anode current and 2·5mA standing screen current in each valve. Some patience is well worthwhile to trim the X-amplifier to optimum linearity by applying these measures. The Y-amplifier output-stage, V2 and V3, can be treated analogously if required, though—because one generally uses less Y-deflection—matters are here much less critical.

Note that normally pins 7 and 11 of the c.r.t. are the Y-plates and 8 and 10 the X-plates; the functions have been changed over here and the c.r.t. turned through 90° to compensate. This has two advantages for the present design. Firstly, it uses the plates nearer the final anode, and thus with higher deflection sensitivity, as X-plates, reducing the total deflection amplitude required from the critical X-amplifier. This measure was decisive in enabling success to be achieved without resort to a mains transformer with higher voltages than a "normal" h.t. winding of 350V. The second advantage is that it brings the clamping-plate of the standard mount and mumetal-can unit horizontal instead of vertical as in the original government apparatus. This allows a mounting-arrangement which is very simple to carry out and needs a minimum of tools and no unusual parts.

The X-amplifier can be fed from any desired external signal at SK3 with VR4 as gain-control. The final position of S1, "External-X", brings this function into operation and mutes the timebase. The X-amplifier then has an effective maximum sensitivity of about 250mV r.m.s. for full-screen deflection, and a level response over just the normal hi-fi *audio* range. The X-amplifier in this function is neither capacitively nor inductively compensated, unlike the Y-amplifier, and has just the simple volume control VR4 as signal-attenuator, which is therefore *roughly* calibrated in volts per centimeter. This is satisfactory, because the expense of full duplication of the

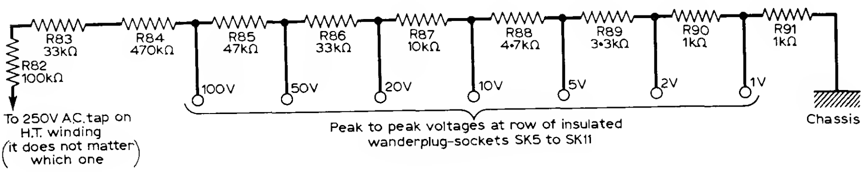


Fig. 9. The circuit of the Y calibrator. Resistor R82 may need reduction or omission, as one half-wave is flattened at the 250V transformer tap, on the conduction half-cycle of the main h.t. rectifier on this side. This reduces the operative peak-to-peak voltage.

high-class video-amplifier (Y-amplifier) would be unwarranted for X-deflection, as such a refinement is seldom needed for normal TV experiments.

Three important measures constitute the compensation of the Y-amplifier (Fig. 1) over video-frequency ranges: firstly, the design of the probe; secondly, the adjustment of the five-key attenuator; thirdly, the stray-capacity compensation of the anode-circuits with inductances L1 to L3. It is necessary to state at this point that, on the most sensitive range of the key-set (key 1 depressed) the circuit is unstable for all *inductive* loads connected direct to P1 if the inductance is tuned to the medium-wave range. This is because V1 then functions as a tuned-anode tuned-grid oscillator with the external inductance and L1.

Instability is absent on *all* ranges when using the probe, and is absent on ranges with keys 2 to 5 even without the probe. When working on aerials and tuned circuits, or tracing signals through the r.f. and i.f. stages of receivers, the probe must be utilised. This instability is no design-defect, as it occurs only under circumstances under which the unit should not be used.

The Bridge-bank Attenuator

A simple resistive voltage divider is in practice not frequency-independent if the time constant of the stray capacities with the resistance-values involved lies in the region of a period of the frequencies concerned. This is because the two stray capacities in parallel with the two resistors constitute a second (capacitive) voltage divider in parallel with the "intended" (resistive) one, and the dividing ratios may differ. In such cases, the stray-capacity ratio determines the transient (high frequency) division-ratio, and the resistive ratio the d.c. and low-frequency division-ratio. For complex waveforms with high harmonic content, serious waveform-distortion occurs.

To restore frequency-independence it is necessary to make the dividing ratios of the resistive and capacitive dividers equal, by augmenting the stray capacities suitably with deliberately added small capacitors in parallel with the resistors. This amounts, effectively, to balancing the a.c. bridge circuit thereby represented. It is seen in the present design that keys 2 to 5 switch in four independent respective "bridges" with the required attenuation factors, whereas key 1, the most sensitive range, lets the signal "straight through".

The total stray capacitance of V1 grid circuit is across SK1 when key 1 is depressed, but is back-attenuated in all the other settings, so that no exact single effective capacitance is operative at SK1 on all ranges, and the probe-capacities should theoretically be switched too. However, this would be very clumsy, and is in practice not necessary. The probe is successfully made-up as a single fixed-balance bridge (see below), as a result of two mitigating measures. Firstly, omission of a fixed capacitor across R1 compensates errors (a value approximately equal to the settings of TC1 to TC4 would otherwise be needed). The stray grid capacitance of V1 takes the place of such a capacitor, very approximately. Furthermore the probe is fitted with a sufficient length of coaxial cable (about 2yd) to make the capacity of this

cable the dominant contributor to the total input capacitance behind the probe. The percentage contribution of the attenuator is thus small, and variations over the ranges are negligible. The probe is then bridge-balanced once and for all against the capacitance of the coaxial cable.

It need hardly be pointed out in this respect that a definite length of cable must be chosen. It is not possible to insert extension lengths between probe and Videoscope indiscriminately if test-apparatus and Videoscope have to be far apart. The required maximum cable length must be decided upon, and the probe designed for this.

The conditions for balance of an attenuator bridge are that the product of each resistor and its total parallel capacitance (stray and intentional) must be equal for all sections. The smaller resistors must thus be fitted with the larger parallel capacitors. It is only the ratio of the capacities which is important; the absolute values are unimportant for balance. Thus, if the bridge is balanced for a particular combination of total capacities, balance is undisturbed if *all* capacities are increased or decreased by the same *factor* (*not* the same amount!). In practice, one will aim to keep the actual capacities small, to avoid total loading on the signal, yet not so small as to be unmanageable.

It is convenient to make the smaller capacity (across the larger resistor) in the form of an ordinary ceramic trimmer, for final true balance adjustment under working conditions. The advanced theorists among our readers can certainly measure strays with suitable capacity-measuring instruments (e.g. transistorised micro-bridge, or observation of shift of resonant frequency of a tuned circuit using a grid-dip meter), and calculate the correct required capacitors to establish balance, selecting and wiring these then as fixed capacitors in place of TC1 to TC4. However, such measurements are laborious and subject to arithmetical and measuring errors.

A much simpler method of final adjustment is to observe a waveform of *known* shape, and trim for optimum display with minimum distortion. Suitable test-waves are square waves, i.e. waveforms with sudden transients (exciting the capacitive branch) and long steady states (exciting the resistive branch). Fig. 10a shows an ideal square-wave from a signal generator. Fig. 10b shows the resulting display on the c.r.t. if the trimmer TC1 to TC4 in circuit is set at too high a capacity, and Fig. 10c the result when balance has been passed and the trimmer is screwed-out to too low a capacity.

Checks must be made at several basic repetition frequencies over as wide a range as possible.

However, it is not essential to use a square wave; the requirement is merely a wave with a broad harmonic spectrum, i.e. one containing sudden transients as well as "waiting" periods. In this way the bridge-bank attenuator will be known to be aligned prior to construction of the probe, so that new distortion subsequently obtained with the probe in circuit, using exactly the same procedure, can only be due to unbalance of the probe-bridge and not to confused multiple effects.

The reasons for having chosen a piano-key switchbank for the bridge-bank of the prototype Videoscope were in the interests of an open layout, giving easy grouping and access to the trimmers in a row. (A rotary wafer switch would give a clumsy arrangement, and because of the tight packing, cross-strays between the separate bridges would result.) A mutual release mechanism should be chosen, i.e. depression of any key releases any other one already depressed. If this component is difficult to obtain, the same function is obviously realiseable with a row of five ordinary d.p.d.t. toggle-switches—but it must be remembered to switch only one down at a time!

The Probe (Fig. 11)

This is designed to have an attenuation factor of 15. Other factors are usable, e.g. a higher factor is obtained by increasing the resistors and decreasing the capacitors by the same factor. This sacrifices overall gain but further reduces capacitive loading on the signal source.

Several probes of different characteristics can be built if desired, making one with very high resistances and low capacitances for observations on signal sources of high impedance where the slightest capacitive loading is serious.

If such ultra-high impedance probes are to be used with high signal amplitudes or d.c. components of several thousand volts, as in TV sets, it must be remembered that the voltage rating of each capacitor in the probe is generally only about 500V. The number of sections and thus the total length of the probe should therefore be increased, keeping to the same individual values as shown in Fig. 11. The attenuation factor and the maximum input voltage are then both directly proportional to the number of sections used.

It is important to use close-tolerance, accurately matched resistor and capacitor chains for the sections. Balance should first be determined experimentally for the chosen length of coaxial cable between probe and Videoscope, using only a single section R92/C61 and selecting various values for C62 until the balance criteria on test waveforms applied to the prod are satisfied.

The number of sections desired can then be added by copying the component values of the balanced section as closely as possible.

E.H.T. Probe

If an e.h.t. probe is required this can be about 18in long and have 40 sections so that it will accept a.c. or d.c. components of up to 20kV at the prod with a total resistance of about 200MΩ.

The resulting maximum deflection sensitivity on the Videoscope will still be some 5V/cm, which is more than ample, and the maximum signal waveform amplitude displayable on the c.r.t. is about 20kV. Measurements can

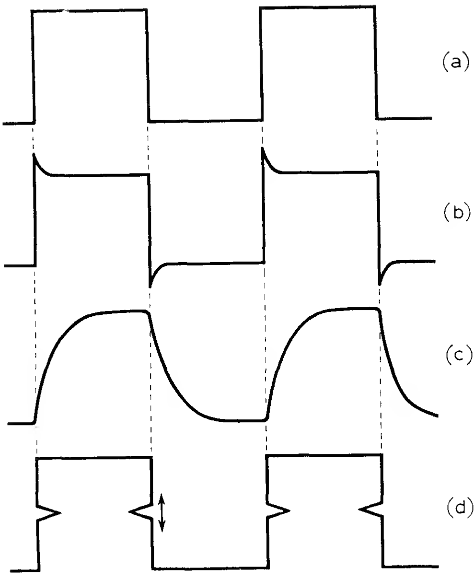


Fig. 10. The waveforms referred to in the text: (a) the ideal square wave; (b) the distorted square wave obtained when the bridge-bank trimmer capacitor is set to too large a capacity, and (c) when set to too small a capacity; (d) sync injection spikes on the square wave when the sync-setting is too intense, or when the floating screen between V4 and V6 is not properly adjusted.

thus be made directly on the e.h.t. pulse winding, rectifier and smoothing capacitor.

A further great advantage is the facility to make an accurate measurement of the d.c. e.h.t. voltage present on the picture tube final anode or elsewhere. The attenuation factors are known and the deflection sensitivity set is also known and the d.c. e.h.t. component is observed as square pulses of "correct" amplitude in relation to calibrations when tapping the prod of the probe in rapid succession on to the e.h.t. points to be tested.

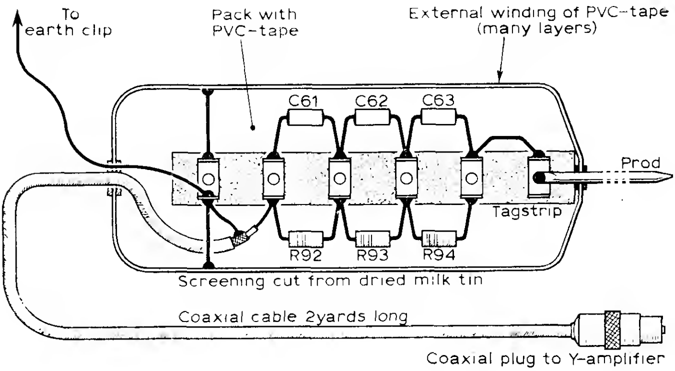


Fig. 11. The construction of the general purpose Y-amplifier probe.

A word of warning: *only experienced constructors with good understanding and ability to work neatly should build or use an e.h.t. probe.* A stout paxolin tube of correct length and generous diameter should be obtained and the attenuator sections built on to a long length of tag-strip as shown in Fig. 11.

This should be held centrally in the paxolin tube like the wick of a candle and the whole tube filled with transformer wax or other approved insulating compound (not candle wax). After allowing to cool and set hard, the outside of the tube may be lapped with tinfoil as screening, securely connected to the outer braiding of the coaxial cable.

A second paxolin tube of larger diameter should then be pushed over the whole lot and the new annular space again poured full of molten insulating compound. Make sure that the prod end of the tinfoil lapping has adequate clearance from the prod (about 2in) and is well below the surface of the insulation compound to avoid e.h.t. flash-over.

It is dangerous to use an e.h.t. probe of poor or indifferent insulation. When using the probe, hold it well back from the prod-end.

General Purpose Probe

The three-section probe of Fig. 11 leads to a small, handy unit which is a good compromise for general use. It will tolerate and display a.c. and d.c. inputs up to about 1·5kV, enabling most measurements in TV sets and elsewhere to be performed with safety. The calibration of the Y-amplifier controls of the Videoscope should be made correct with respect to this probe. Any other special probes made in addition should then be labelled with the *additional* attenuation or gain factor resulting with respect to the Y-amplifier calibrations.

The Anode Peaking Coils (see Fig. 1)

Any ordinary resistance-capacity coupled amplifier has an upper limit set to the operational frequency range due to the stray anode and grid capacitances causing increasing shunting of the anode loads at higher frequencies.

Well-designed amplifiers of this type without special compensation, apart from simple negative feedback, can reach upper frequency limits of several hundred kc/s with usable gain.

This is not sufficient for the Videoscope, so that additional measures have been adopted to raise the upper frequency limit to about 2Mc/s. Regarding the low-frequency end, there is little to be said—this is merely a question of using coupling capacitors of adequate size in R.C.-coupled amplifiers.

The principle of anode peaking-coils inserted at the h.t. end of the anode load resistors is to give a resonance at or near the otherwise resulting cut-off frequency. This is achieved with the coil and the stray capacitances acting in parallel, representing then, as always in a parallel-tuned circuit, a relatively high impedance at resonance.

The anode resistor is effectively in series with the coil for this tuned circuit, giving tremendous damping of the resonance. This is desirable because the resonance is required to be so broad that a smooth and matched lift to level response is obtained without preference for the resonant frequency or any frequency close to it.

The design must thus be for Q values less than unity. At frequencies well below the resonance, the coil has negligible impedance and is thus as good as absent, i.e. the performance of the amplifier is not disturbed.

Provided that the circuitry and layout as published are *strictly* adhered to, the quoted inductance values will be found satisfactory. L2 and L3 must be matched. The required inductance values can be trimmed experimentally on small formers and using thin enamelled copper wire, checking resonant frequencies with a known parallel capacitor (large enough to swamp self-capacitance), using a grid-dip meter.

As a rough guide, about 300 turns of the type of thin enamelled wire taken off the primary of an ordinary wireless output transformer, pile-wound on to one of the standard small v.h.f. coil formers with screwed slugs (retain the slugs), should be required for L1 and about 400 turns for L2 and L3.

These windings will normally make the required inductance lie within the range of the slug. Further variation, to avoid rewinds unless really necessary, can be got by trying various slugs (v.h.f. types or medium wave types).

After dipping each finished coil in insulating lacquer and allowing to dry, a good layer of p.v.c. tape should be wound on and then a single lap of tin-foil as screening. This should be insulated with an interposition of p.v.c. tape at the overlap in the same way as the screen winding of a mains transformer to prevent representation of a shorted turn. A lead from the screen should go to the nearest tag when wiring up.

The outside of the whole screened coil should be lapped with further p.v.c. tape and finally dipped in insulating varnish. Stiff connecting leads should have been fitted and anchored so that the overall final appearance is that of a paper capacitor in roll form.

For those readers wishing to be more exact and possessing the necessary instruments the exact coils required can be determined by the same procedure used by the author in the prototype in establishing the initial design.

Two simple formulae are used for this. If Ra is the operative load resistor in the anode circuit and Ca is the total stray anode capacitance then the product of Ra (expressed in MΩ) and Ca (expressed in μF) gives the cut-off time-constant in seconds.

The reciprocal of this is $2\pi f$c where fc is the nominal cut-off frequency without compensation. It is the frequency for which the amplification has fallen to $1/\sqrt{2}$ of the full value. We must now choose a value for the coil inductance such that the ratio of coil impedance to anode load resistance at fc is given by $(\sqrt{2}-1)$.

Then the gain is lifted back to level at fc instead of being about 30% down and remains level to beyond the resulting resonant frequency which lies at $\sqrt{2}$fc. Combining the formulae we get the single requirement for choice of inductance for the compensation coil:

$$L = 0{\cdot}414 . Ca . Ra^2$$

L = required inductance (μH)

Ra = anode load resistance (kΩ)

Ca = total anode stray capacitance (pF).

Here the only unknown is the anode stray capacitance, which must be measured. This is best done with a grid-dip meter by comparing the resonant frequency of any available test coil (even a makeshift multiple loop of connecting wire giving resonances in the short-wave band is suitable) wired between valve anode and chassis with the resonant frequency of *the same* coil wired across various known or calibrated variable capacitors.

The inductance of the test coil need not be known for such measurements. The anode load resistor should be disconnected at the h.t. end for these measurements with a test coil.

Do not forget to have the valve inserted in the socket and the screening can in place, and all wiring complete at the anode, so as to measure the *total* operative stray capacity. The valve heater should preferably be operating but there should be no h.t. applied to anode or screen.

It is important to point out in conclusion that the same procedure for determining the optimum peaking inductance can be applied to video stages of TV receivers. Many good sets already incorporate such coils. The question is of definite topical interest for the constructor at present because of the need to raise the bandwidth of all stages when converting older BBC standard receivers to dual-standard operation.

Better C.R.T.

The Videoscope was primarily devised for using the cheap surplus c.r.t. VCR139A. This tube has several shortcomings, the circuit in Figs. 1, 2 and 3 being specially matched to these.

Improvements in performance are possible with a more expensive commercial tube. The Mullard DG7-32/01 is eminently suitable for the Videoscope; it costs about five pounds but brings many advantages to the circuit. It is nearly the same size as the VCR139A, so no major structural changes are required, and the operating voltages are so similar to those provided for the VCR139A that only minor changes of some component values are needed. Apart from its superior brilliance and focus, the DG7-32/01 has about 50% more Y-deflection sensitivity than the VCR139A, so that appropriate reductions in the anode load resistor values in the Y-amplifier become possible, increasing the *level* response up to some 1·7Mc/s and the range of usable gain up to 3Mc/s, the lower limit remaining at about 5c/s as previously.

It is the purpose of this section to sketch the necessary alterations involved in this conversion and to give details of modifications to the sync-circuits for obtaining improved lock for r.f. signals in keeping with the increased bandwidth obtainable. Whilst maintaining adequate sync rigidity down through the audio range and right down to well below the power mains frequency, the modified sync-circuit still gives rigid lock on a sinewave signal at 3Mc/s fed into the Y-amplifier from an ordinary r.f. signal generator, even with the otherwise unmodified Videoscope still using the VCR139A tube.

Increasing the Bandwidth of the Sync-circuits (see Fig. 12)

We will treat these modifications first, as they may even be useful if keeping the VCR139A tube when waveforms at particularly high frequencies are to

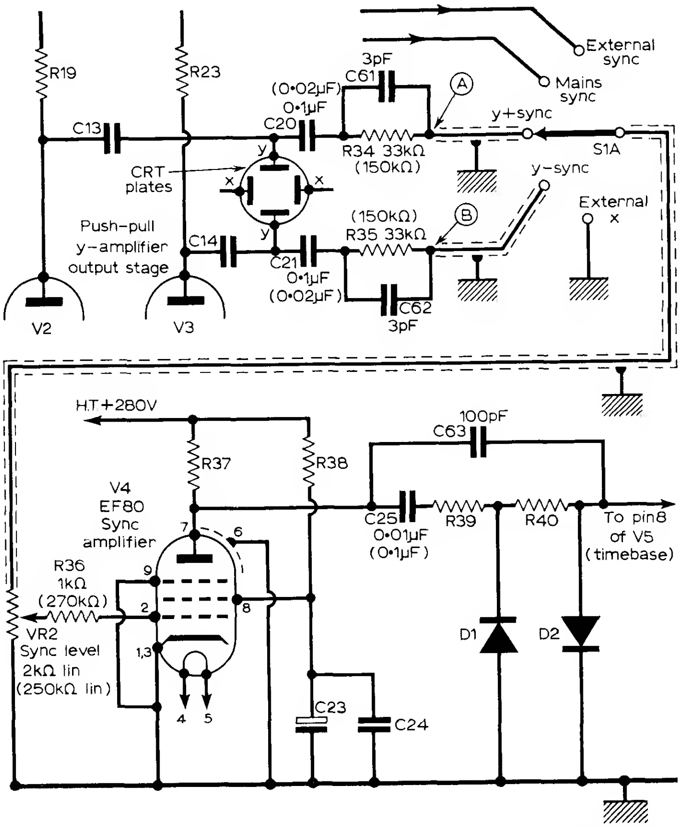


Fig. 12. Modifications to the sync circuits to obtain uniform sync action for audio and radio frequencies. Bracketed component values are those given in the original circuit (Fig. 2).

be 'scoped. The following simple changes of values are required. C20 and C21 should be changed in value to 0·1μF each. R34 and R35 should be reduced in value to 33kΩ each, and a 3pF ceramic, 500V capacitor should be wired in parallel with each of these resistors. The sync-level potentiometer VR2 should be exchanged for a linear 2kΩ potentiometer and R36 reduced to 1kΩ (see Fig. 12).

In the anode circuit of sync amplifier V4, C25 should be reduced to 0·01μF and a 100pF 500V tubular ceramic capacitor bridged from one end of R39 to the far end of R40 (i.e. effectively between V4 pin 7 and V5 pin 8).

The modified circuit gives excellent and uniform sync of all signal frequencies from the lowest audio right up to several Mc/s, but somewhat less effective than the audio frequency sync obtained with the original circuit which, for its part, does not work above the highest timebase frequency provided, i.e. above about 50kc/s.

Increasing the Bandwidth of the Y-Amplifier

The original circuit (Fig. 1) has a *level* Y-gain from 5c/s to 1Mc/s and usable gain up to over 2Mc/s. The following modifications to the Y-amplifier, if used with the DG7-32/01, raise these upper-frequency limits by about 50% for the same effective sensitivity in V/cm as previously obtained.

The component changes are: Reduce R9 and R23 each to 6·8kΩ 1W and L2 and L3 each to 575μH. Increase R43 to 4·7kΩ ½W.

Modifications to the C.R.T.-Network for using the DG7-32 Tube

Fig. 13 shows the optimum circuit for operating the DG7-32/01, a high-intensity low-voltage tube in which the cathode and final anode currents are high. Four zener diodes, D11 to D14, are used to stabilize the e.h.t. bleeder potentials against variation with brilliance setting.

D11 and D12 replace R73 and R74 to keep the shift control voltages independent of the brilliance setting. D13 and D14 maintain a constant voltage across VR8, the brilliance control, and therefore reduce the extent of interaction of focus and brilliance controls. R104 has been added to maintain sufficient standing current through D13 and D14. All four zener diodes are type ZL56.

Such stabilisation was not found to be required for the VCR139A but it can be tried if some specimens of this tube should draw higher anode current.

To cater for the higher total current of the DG7-32/01, R72 and R69 are reduced to 2·7kΩ. This leads to somewhat poorer smoothing on both positive and negative e.h.t. feeds, and the disposition of decoupling and smoothing capacitors around the shift controls had to be altered as shown in Fig. 13. This is important, since the DG7-32/01 is marked by particularly fine trace focus in conjunction with high trace intensity, requiring good e.h.t. smoothing.

Finally, to bring the point of true focus into about the centre of the track

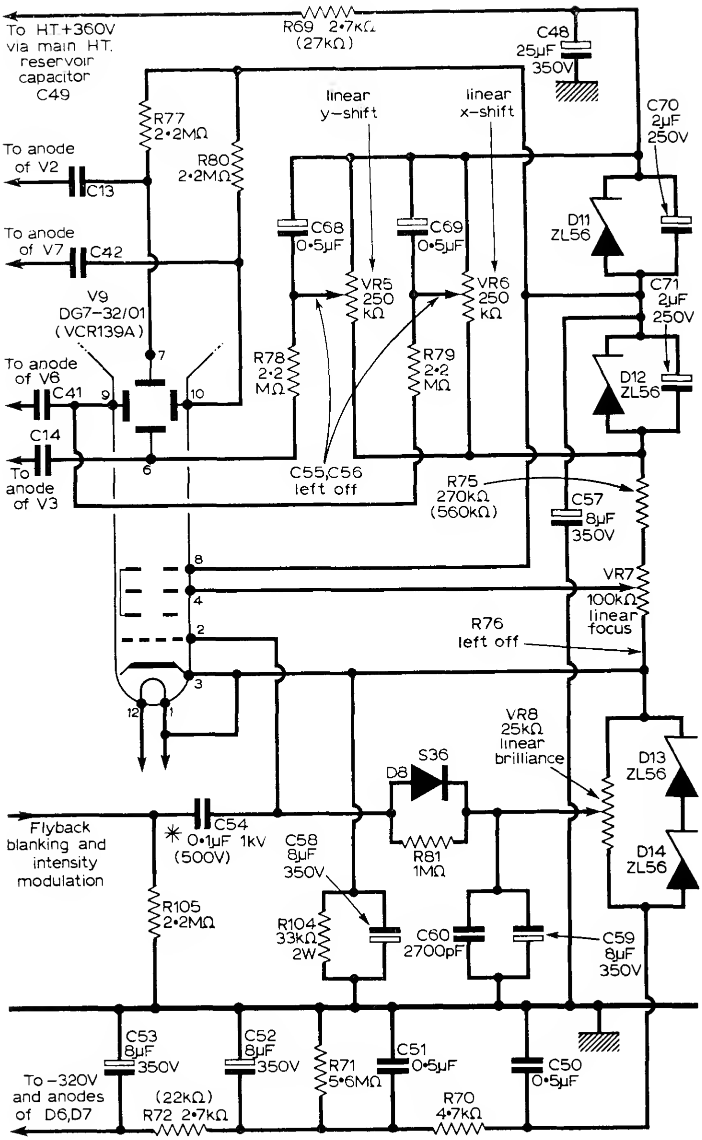


Fig. 13. A stabilised e.h.t. network and optimum circuit for operating the Mullard DG7-32/01 c.r.t. in the Videoscope. It is important to change the heater tapping on the mains transformer to 6·3V from the original 4V. C54 should have the higher working-voltage when used with an intensity-modulation amplifier.

of VR7, R75 had to be reduced to 270kΩ in this circuit, and R76 at the bottom end of VR7 had to be discarded.

This circuit permits operation of the DG7-32/01 up to a very high trace intensity without undue loss of focus so that a standard TV picture raster may be built-up at sufficient intensity and with adequate focus for displaying a C.C. TV-picture for setting-up purposes (viewfinder for a C.C. TV camera) in addition to normal oscilloscope work.

If TV-picture display is not required, and if one is satisfied with moderate trace intensity for normal oscilloscope work, a simpler circuit can be used. Compared with the original circuit this involves only the following modifications: Shunt R73 with a 56V zener diode and increase VR8 to 100kΩ (linear), shunting the new potentiometer with a 68V zener diode. The respective polarities of the diodes are the same as in Fig. 2. Reduce R75 to 470kΩ.

Structural Changes in C.R.T. Circuit

The DG7-32/01 is about an inch shorter than the VCR139A, but the dimensions at the screen-end are virtually identical. Thus no changes are required in the hole and mask arrangement on the front panel, but the whole assembly must be moved forwards about 0·8in to 1·0in, according to the exact thickness of the c.r.t. base.

If the base-plate on the end of the mumetal-screen assembly of the VCR139A is removed and a new plate cut from sheet aluminium, mounting this on to the mumetal screen with the same three rocker screws and bolting the base for the DG7-32/01 on to the new plate in the correct orientation (Fig. 14), it will be found that the new c.r.t. fits exactly. No changes are required to the chassis-fixture bracket except to drill three holes in the bottom flange in order to move the whole assembly about an inch forward.

It is advisable to elongate the new base-plate upwards as shown, mounting an 8-way tag-strip along the top. The new c.r.t. network of Fig. 13 can then be wired up as shown in Fig. 14, although slight departures in layout are tolerable.

Timebase Range

The X-sensitivity of the DG7-32/01 in the circuit (Fig. 13) is identical to the original circuit with the VCR139A where the Y-plates were used for X-deflection, so that no changes to the timebase circuit or its calibration are required. For the same reasons, no changes are required to the X-amplifier or its calibrations.

Modifications to the Flyback-blanking Circuits

The DG7-32/01 requires a greater flyback-blanking amplitude than the VCR139A, and the most convenient method of achieving this is to place a suitable blanking amplifier between the junction of R52, R51 and the line

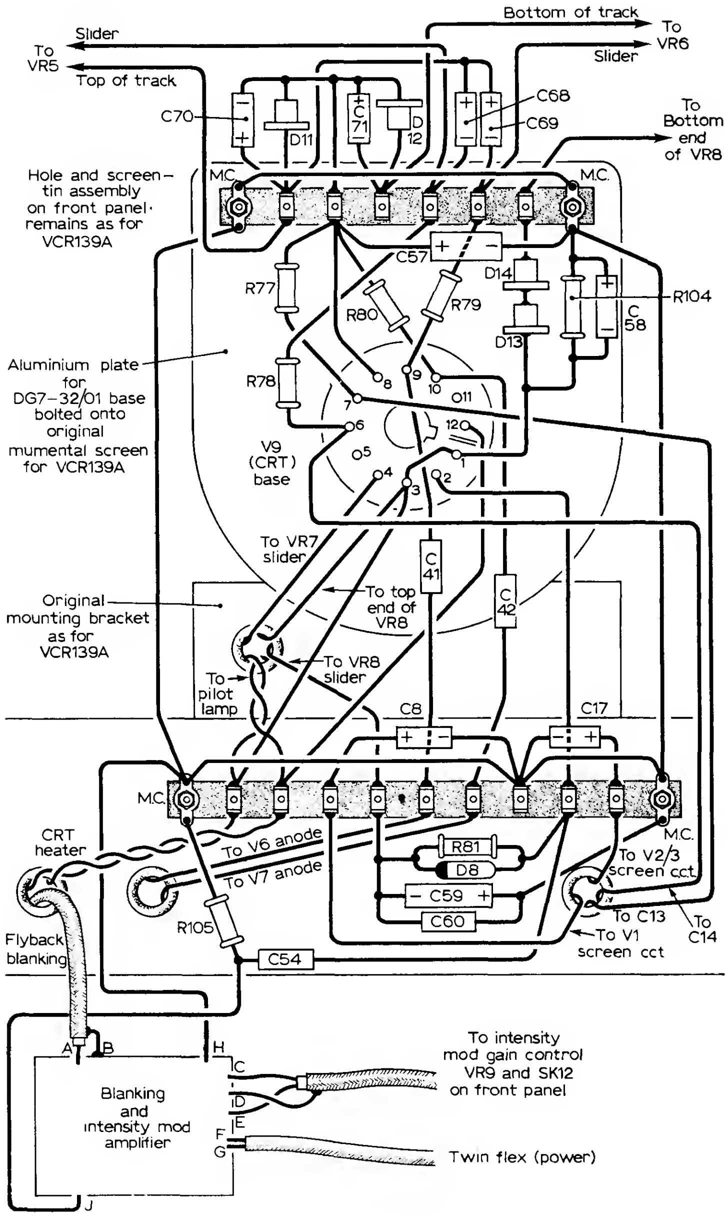


Fig. 14. Connections for the DG7-32/01. The physical details of the hole and screen-mask assembly remain as for the VCR139A, but the mounting-bracket is moved forward by 0·8in to 1·0in.

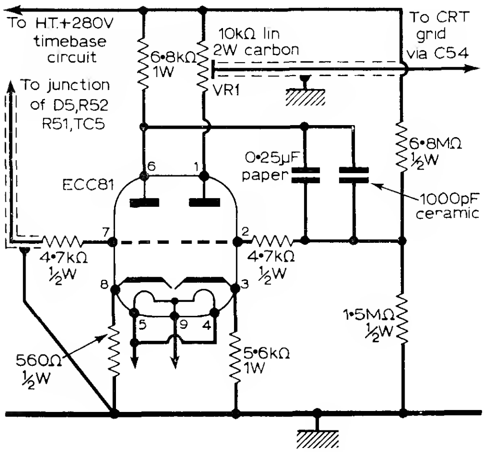


Fig. 15. A circuit suitable for a flyback-blanking amplifier when other *c.r.t.s are used which require* greater *blanking drive. It must be checked that the mains transformer has adequate capacity to provide the additional 0·3A at 6·3V, otherwise an additional subminiature 0·3A heater transformer will be required. R54 should have a similar resistor wired in parallel when the blanking-amplifier is used.*

to the c.r.t. grid. Fig. 15 shows a suitable circuit which can be wired-up on a miniature bracket-chassis screwed to the wooden cabinet base behind the main chassis and beside the c.r.t. network. The pre-set potentiometer VR1 in Fig. 15, should be set for smooth flyback blanking at all timebase speeds.

General Remarks

Although these notes are intended for those whose primary interest is not cost but in performance similar to the average portable TV-servicing oscilloscope of commercial manufacture, those with a smaller purse should nevertheless not be discouraged from building the Videoscope around the surplus VCR139A. Performance is adequate for most ordinary TV servicing, and indeed the author used the VCR139A version very successfully for closed-circuit television design work, *before* embarking on the DG7-32/01 conversion.

The original X-amplifier gives very good performance up to 1Mc/s (with X-gain full up). A sensitive test satisfactorily passed was to connect the X-amplifier input and the Y-amplifier input (via the probe) each to respective medium-wave tuned circuits both excited by being placed close to a grid-dip meter running at 1Mc/s. The two tuned circuits were tuned off a little in opposite directions so that the signals were 90° out of phase with each other. An almost perfect circle of more than an inch in diameter could be traced on the c.r.t. screen.

The geometric accuracy of such a circular trace is a sensitive test for any distortion in the 'scope (or signal-generator for that matter). Any 'scope which will "write the 1Mc/s circle" is capable of very useful work on amateur television equipment and domestic TV receivers, whereas further bandwidth in the Y-amplifier such as is provided in the modified version using the DG7-32/01 tube is very useful for special work and more detailed observation of pulse-flanks in television and other circuitry.

A Y-bandwidth greater than 5Mc/s, which the modified Videoscope will approach, is hardly ever required for normal television work, as higher-frequency transients cannot be accommodated in the normal TV waveform anyway. But for special C.C. TV or colour systems using increased numbers of lines per frame, bandwidths of up to 20Mc/s or more can be very useful.

Experiments showed that the question of incorporating a flyback blanking amplifier according to Fig. 15 is a borderline case for the DG7-32/01. Using the original "straight-through" blanking circuit without an amplifier, as described for the VCR139A, blanking action was equally satisfactory with the DG7-32/01 up to the levels of trace brightness previously also obtainable with the VCR139A.

If the brilliance is then turned up to the much higher levels which the DG7-32/01 is capable of, the flyback reappears and blanking is therewith incomplete.

The more advanced amplifier circuit of Fig. 16, for which a detailed layout diagram is given in Fig. 17, was the result of efforts to devise further functions which such an amplifier could also assume within the Videoscope, to justify its inclusion more definitely. An oscilloscope is immediately rendered

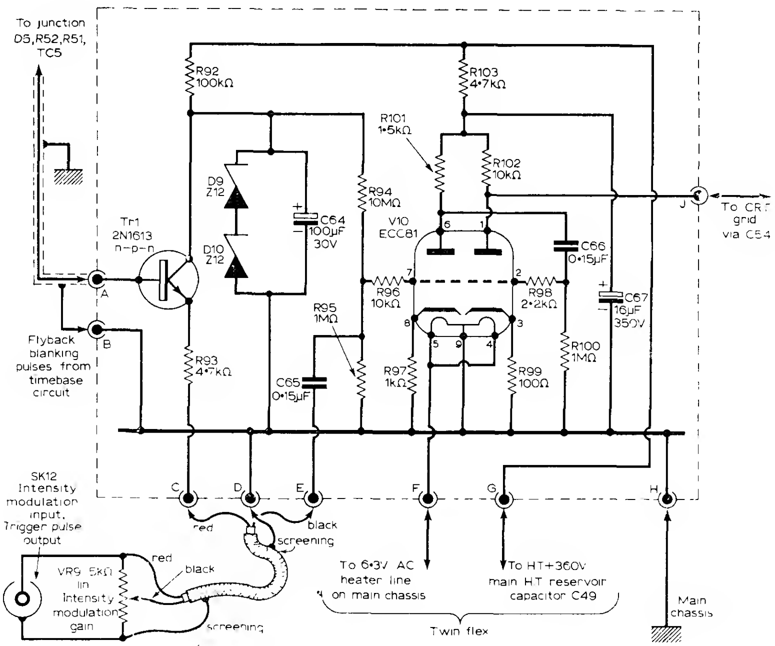


Fig. 16. An advanced flyback-blanking amplifier with manual control of the blanking-level. This is also suitable for general-purpose intensity modulation and video amplification for TV picture display on the Videoscope.

more useful if a third parameter can be displayed in the form of beam intensity modulation at the control grid, in addition to the X and Y deflection signals.

Flyback blanking is one form of beam intensity modulation, and thus it seemed reasonable to undertake experiments to make the amplifier of more general application while also catering for optimum flyback blanking during normal oscilloscope work.

The design of Fig. 16 was developed with due regard to the principal requirements for intensity modulation in a general-purpose TV oscilloscope. These are largely twofold. Firstly, when using the oscilloscope in conjunction with a wobbulator for TV receiver alignment, it is useful to inject frequency markers by way of intensity modulation.

Secondly, intensity modulation is required to display a TV picture in the conventional manner. The latter function dictated the design bandwidth of about 1Mc/s for the amplifier of Fig. 16 (this is sufficient for obtaining a clear TV picture on such a small screen).

Design Principles

For general applications, an input socket for an external intensity-modulation signal and a suitable gain control (SK12 and VR9) were provided on the front panel. These had to be at very low impedance, to permit connection to the amplifier sub-chassis via screened cable. Twin screened cable is better than two separate screened cables, since the self capacitance between the two cores then acts in a response-correcting sense.

The transistor Tr1 is the simplest and most convenient matching device. It operates as emitter follower, transforming the high-impedance flyback blanking pulses arriving from the junction of R51/R52/D5 (main chassis) to the required low impedance level. These low-impedance pulses may be taken off at SK12 if required, this constituting a most useful output signal of the Videoscope.

The output is an asymmetrical square wave at SK12 whose negative part coincides with each timebase flyback and whose positive part coincides with each timebase run. Apart from uses as a general-purpose audio-pulse signal generator, this output can be used to trigger or slave-drive all manner of ancillary units such as double-beam switches, wobbulators etc.

The low-impedance flyback pulses from Tr1 are also applied across the track of VR9 and any desired portion of their amplitude can be taken off at the slider and applied to the actual intensity modulation amplifier V10 which is basically similar to Fig. 15. The flyback blanking can therefore be set to optimum with VR9 at all times.

Correct procedure is to turn VR9 to zero, set desired trace brilliance and lock the waveform being observed, and then to advance VR9 until the flyback trace has just disappeared. Since D8 (Fig. 13) restores the d.c. component of the amplified blanking-waveform such that the positive level is clamped to the potential at the slider of the brilliance control VR8, the

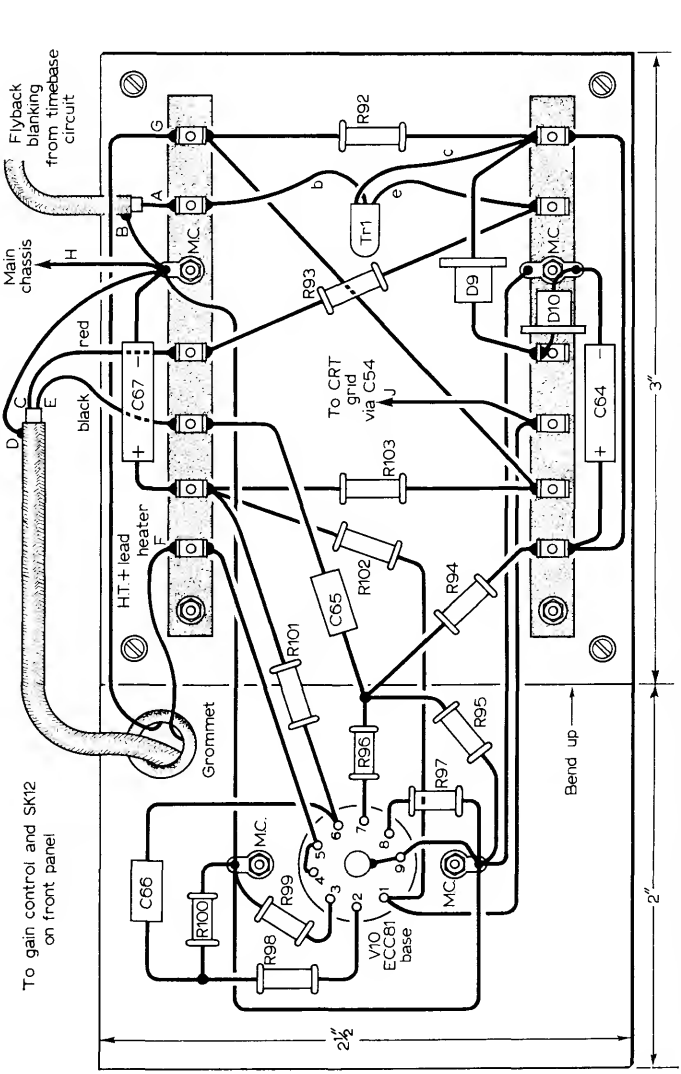


Fig. 17. Constructional details of the advanced flyback-banking and intensity-modulation amplifier. This sub-assembly may conveniently be screwed to the base of the cabinet at the centre of the rear of the main chassis.

intensity-modulation control VR9 has no effect whatsoever upon the brilliance of the timebase trace, influencing the flyback alone.

Two stages of amplification are used in Fig. 15 and in V10 of Fig. 16 in order to preserve the same output polarity as required. In Fig. 16 the first stage of V10 operates at insignificant gain (anode and cathode loads about equal), to avoid loss of bandwidth at the relatively high impedance grid 7 due to the Miller effect (R96, 10kΩ, is required to prevent parasitics on video transients).

The net impedance at the grid of the second stage is much lower, so that the Miller effect due to the higher anode load and higher gain is here insignificant and bandwidth is maintained.

The value of R101 should be selected such that a standard IV video signal applied to SK12 will just modulate the c.r.t. grid fully when VR9 is turned about two-thirds up. R93 should thereafter be selected such that flyback blanking is just complete at the highest trace brilliance setting usable when VR9 is turned fully up and nothing is connected to SK12. The component values shown in Fig. 16 satisfied these conditions in the prototype.

D9 and D10 are two small 12V zener diodes to establish the positive collector voltage of 24V for Tr1 (*npn* transistor). R92 and C64 give the necessary voltage drop from the h.t. line and the essential very high degree of smoothing, since the same derived 24V supply is used to set the operating point at V10 grid (pin 7) via R94.

Both stages of V10 operate strictly linear in Class A, but Tr1 operates in Class B, being cut-off in the absence of a base signal (during flybacks) and keyed-on during the timebase runs (positive output pulse for duration of timebase run at SK12).

Any other small silicon *npn* transistor (*not* germanium and *not pnp* type!) with a voltage rating of 30 or more and a current gain of at least thirty may be substituted for Tr1, regardless of make.

Television Picture Display

The following remarks apply to the display of a television picture on the Videoscope fitted with a DG7-32/01 tube, picked up by any c.c. TV camera operating with an EMI 10667 vidicon tube (or direct equivalent), which is the most common camera tube at present to be found in amateur c.c. TV equipment.

The Videoscope fitted with a DG7-32/01 operating in the circuit of Fig. 13 and fitted with the intensity-modulation amplifier shown in Fig. 16 is eminently suitable for displaying a small TV picture from such cameras for setting-up purposes and "viewfinding", in addition to normal oscilloscope work on the c.c. TV equipment. The connections for TV picture display under these conditions are as follows:

1. Connect the Videoscope Y-amplifier input in parallel with the vidicon field-scan coil, *without* using the probe. Depress the "25V/cm" course

gain key, when correct field amplitude will be found about midway on the fine control of Y-amplifier gain.

2. Connect the Videoscope External-Sync input in parallel with the vidicon line-scan coil, switch X-function switch to "external sync" and set timebase coarse speed to fastest range. Correct line lock for both 405 and 605 lines will then come within the fine control range.
3. Connect the video monitor output signal (standard 1V p.p. positive video signal) from the c.c. TV camera to the new intensity-modulation input socket SK12 on the Videoscope and advance the gain control VR9 as required.

Picture quality is surprisingly good, considering the circumstances. Individual scale divisions on a 2in moving coil meter televised via the vidicon camera in this manner were clearly visible and the pointer reading could be read-off accurately. Line lock was extremely secure even when the sync level control was advanced only about quarter-track.

Neither 405 nor 625 lines can be resolved individually, so that the image is continuous without line structure. The general appearance is that of the type of snapshot customary from the cheaper type of box-camera and thus is fully adequate for setting-up and viewfinding.

Variations of Procedure

If the cable from the vidicon field-scan coil is unplugged from the Y amplifier and the video cable is transferred from the intensity modulation input to the Y-amplifier input, the display on the c.r.t. screen immediately reverts to the normal line-video oscillogram *without changing the settings of any controls.*

It is thus possible to change back and forth in a couple of seconds between the two most important c.c. TV monitoring settings of picture display on the one hand and line-video oscillogram on the other hand. The resolution of the latter is excellent with the DG7-32/01.

With the same 2in moving coil meter being televised, movement of the thin pointer was seen as a sharp pulse riding across the video oscillogram and with a little practice it was possible to read-off the pointer indication with reasonable accuracy that way too!

The video oscillogram is most useful in studying correct adjustment of beam and target controls for the vidicon, particularly in relation to illumination and required lens-stop. It also permits correct adjustment of sync-to-vision ratio. All these adjustments are less clear on a picture display, which then serves the principal purpose of "viewfinding" once the camera controls are set correctly.

A second variation of procedure is dictated when a picture display is desired under conditions of high ambient illumination. The cable from the vidicon line-scan coil can be unplugged from the "external sync" input of the Videoscope and plugged into the "external X signal" input, switching

the X-function over to mute the Videoscope timebase and operate the X-amplifier off an external signal.

The line scan signal for the vidicon is almost a pure square-wave with a tiny sawtooth component. The latter is amplified by the Videoscope X-amplifier to give a small line scan on the DG7–32/01, the entire picture then being pushed to the right on account of the large square-wave component. The latter can be cancelled with the X-shift control. After suitable reduction of the Y-gain, any desired degree of reduction of picture size can be achieved by suitable setting of X-gain, with corresponding increase of brightness. In this manner an intensely brilliant picture about half an inch in diameter can be produced with sufficient clarity for general "viewfinding" in aiming the c.c. TV camera in a strongly lit room.

It is not possible to increase the picture diagonal above about three-quarters of an inch in this arrangement, since line scan becomes grossly non-linear thereafter due to overloading of the X-amplifier on account of the large square-wave component of the vidicon line-scan waveform, but the whole point of this setting is to get *small* pictures for conditions of high ambient illumination.

If the maximum picture size (2in diagonal) is required, the first setting described, with the vidicon line-scan merely used to synchronise the internal timebase, is essential. Linearity of both line and field is then excellent.

Black Level

The reference level of a television video waveform is given by the sync pulses, so that correct black level clamping is ideally given by a positive d.c. restorer diode. This is exactly the opposite requirement to flyback blanking at the grid of the c.r.t. during oscilloscope display, which needs a negative d.c. restorer (D8, Fig. 13).

If D8 in Fig. 13 is reversed, to constitute the positive d.c. restorer for TV picture display, the functions of brilliance and flyback blanking controls become strange and unwieldy during oscilloscopy. The brilliance control then has to be set to a threshold value and actual trace brilliance adjusted with the blanking control, which is most confusing.

Thus it is essential to maintain the polarity of D8 as shown in Fig. 13, since only then do brilliance and blanking controls operate correctly and independently during oscilloscopy. If D8 were to be omitted altogether, the settings of brilliance and blanking would both no longer be independent of timebase speed.

D8 *must* therefore be of incorrect polarity for TV picture black level clamping and consequently clamps the picture highlights, at whatever level they may lie, to the brilliance level selected at the slider of the brilliance control VR8. Thus the picture gets weaker and brighter as VR9 is backed off, instead of weaker and darker as is more familiar with normal TV receivers when contrast is reduced.

However, this is not confusing and therefore is not considered as being a

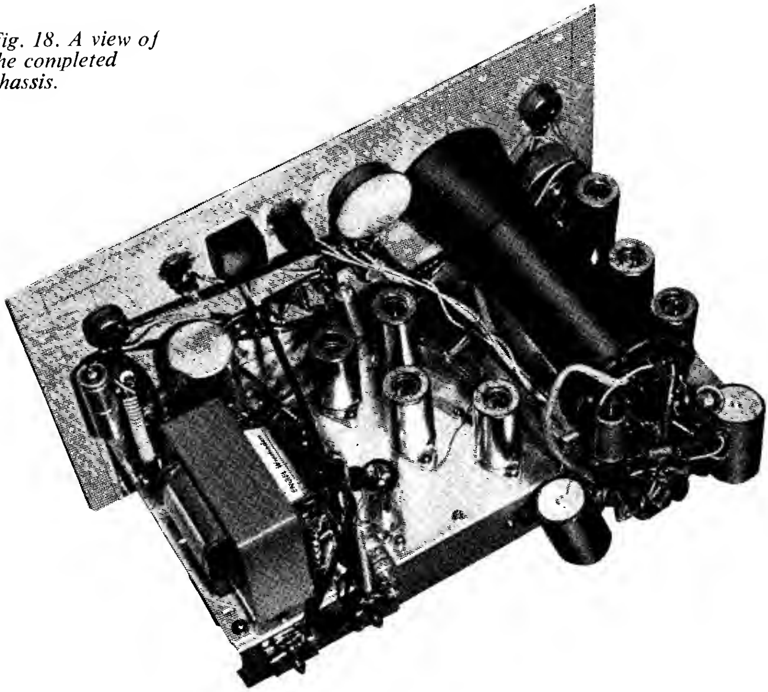


Fig. 18. A view of the completed chassis.

disadvantage. After all, oscilloscopy is the primary function of the Videoscope and not high-quality TV picture display. Indeed, settings of the brilliance and intensity modulation controls were found with both polarities of D8 giving identical picture display qualities.

There is in principle no objection to using a circuit such as the one shown in Fig. 16 for TV picture display on any other suitable oscilloscope. The only requirements are sufficient beam intensity without undue loss of focus and accessibility of the c.r.t. grid. The facilities of external sync input for the built-in timebase and a Y-amplifier will be found on all oscilloscopes.

List of Components

Resistors:

R1	1MΩ $\frac{1}{2}$W	R8	1MΩ $\frac{1}{2}$W
R2	820kΩ $\frac{1}{2}$W	R9	4·7kΩ $\frac{1}{2}$W
R3	180kΩ $\frac{1}{2}$W	R10	470Ω $\frac{1}{2}$W
R4	1MΩ $\frac{1}{2}$W	R11	1MΩ $\frac{1}{2}$W
R5	38kΩ $\frac{1}{2}$W	R12	270Ω $\frac{1}{2}$W
R6	1MΩ $\frac{1}{2}$W	R13	470Ω $\frac{1}{2}$W
R7	10kΩ $\frac{1}{2}$W	R14	10kΩ 2W
		R15	4·7kΩ $\frac{1}{2}$W

List of Components (*contd.*)

R16 10kΩ 2W
R17 1MΩ ½W
R18 470Ω ½W
R19 10kΩ 2W
R20 22kΩ 1W
R21 68kΩ ½W
R22 68kΩ ½W
R23 10kΩ 2W
R24 100Ω ½W
R25 100Ω ½W
R26 470Ω ½W
R27 270Ω ½W
R28 270Ω ½W
R29 1MΩ ½W
R30 5kΩ 5W w-w
R31 2·2MΩ ½W
R32 2·2MΩ ½W
R33 10kΩ ½W
R34 150kΩ ½W
R35 150kΩ ½W
R36 270kΩ ½W
R37 270kΩ ½W
R38 470kΩ ½W
R39 220kΩ ½W
R40 220kΩ ½W
R41 47kΩ 1W
R42 100kΩ 1W
R43 2·7kΩ ½W
R44 150kΩ 1W
R45 470Ω ½W
R46 3·3kΩ ½W
R47 100kΩ 1W
R48 100kΩ ½W
R49 47kΩ 1W
R50 68kΩ ½W
R51 1MΩ ½W
R52 150kΩ ½W
R53 2·2MΩ ½W
R54 10kΩ 1W
R55 33kΩ ½W
R56 470kΩ ½W
R57 470Ω ½W
R58 10kΩ 2W
R59 68kΩ ½W
R60 68kΩ ½W
R61 10kΩ 2W
R62 5kΩ 5W w-w
R63 22kΩ 1W
R64 100Ω ½W
R65 100Ω ½W
R66 470Ω ½W
R67 47Ω ½W
R68 470kΩ ½W
R69 27kΩ ½W
R70 4·7kΩ ½W
R71 5·6MΩ 1W
R72 22kΩ 1W
R73 270kΩ 1W
R74 270kΩ 1W
R75 560kΩ 1W
R76 47kΩ 1W
R77 2·2MΩ ½W
R78 2·2MΩ ½W
R79 2·2MΩ ½W
R80 2·2MΩ ½W
R81 1MΩ ½W
R82 100kΩ 1W
R83 33kΩ 1W
R84 470kΩ 1W
R85 47kΩ 1W
R86 33kΩ 1W
R87 10kΩ ½W
R88 4·7kΩ ½W
R89 3·3kΩ ½W
R90 1kΩ ½W
R91 1kΩ ½W
All ±10% carbon unless otherwise stated

Potentiometers:
VR1 2kΩ 2W lin.
VR2 256kΩ lin.
VR3 2MΩ lin.
VR4 250kΩ log.
VR5 250kΩ lin.
VR6 250kΩ lin.
VR7 100kΩ lin.
VR8 25kΩ lin.

Capacitors:
C1 47pF cm
C2 220pF cm
C3 750pF cm
C4 1500pF cm
C5 0·25μF pt
C6 2200pF pt
C7 2700pF pt
C8 50μF e350/t
C9 32μF e350/t
C10 50μF e450/550/c
C11 2700pF pt
C12 0·15μF pt

LIST OF COMPONENTS (*contd.*)

C13 0·1μF pt
C14 0·1μF pt
C15 0·15μF pt
C16 0·01μF pt
C17 50μF e350/t
C18 2700pF pt
C19 0·02μF pt
C20 0·02μF pt
C21 0·02μF pt
C22 8μF e350/t
C23 8μF e350/t
C24 2700pF pt
C25 0·1μF pt
C26 2700pF pt
C27 8μF e350/t
C28 2700pF pt
C29 8μF e50/t
C30 8μF e350/t
C31 0·1μF pt
C32, 33* 100pF cm
C34, 35* 1000pF cm
C36, 37* 0·01μF pt
C38, 39* 0·1μF pt
C40 0·1μF pt
C41 0·1μF pt
C42 0·1μF pt
C43 16μF e350/t
C44 2700pF pt
C45 0·15μF pt
C46 8μF e350/t
C47 0·01μF pt
C48 25μF e350/t
C49 50μF e450/550/c
C50 0·5μF mp
C51 0·5μF mp
C52 8μF e350/t
C53 8μF e350/t
C54 0·1μF pt
C55 0·1μF pt
C56 0·1μF pt
C57 8μF e350/t
C58 8μF e350/t
C59 8μF e350/t
C60 2700pF pt

* Matched pairs: selected 1:10:100:1000

Code: pt—paper tubular 500V; mp—metallised paper can 500–750V; cm—ceramic or mica 500V; e50—electrolytic 50V; e350—electrolytic 350V; e450/550—electrolytic 450–550V; /t—tubular; /c—can

Valves:
V1 EF80
V2 EF80
V3 EF80
V4 EF80
V5 EF86
V6 EF80
V7 EF80
V8 EZ80
V9 VC139A

Diodes:
D1, 2, 3, 4, 5, 8 S36 fast silicon (Brush Crystal Co., Southampton)
D6, 7 E250 C50 flat selenium rectifiers

Switches:
S1 3-pole 5-way (two wafers each 2-pole 5-way, leaving one pole blank
S2 5-key push button aggregate (piano key unit) each key 2-pole 2-way. See text
S3 2-pole 4-way wafer
S4 Double-pole on/off toggle

Sockets:
SK1, 2, 3 Coaxial sockets
SK4 3-pin mains socket
SK5, 6, 7, 8, 9, 10, 11 Insulated wander sockets

Miscellaneous:
TC1, 2, 3, 4, 5 3–15pF "postage stamp" trimmers
F1 1A slow fuse
LP1 Red, panel-mounting pilot lamp
T1 Mains transformer. Secondaries 350/250/0/250/350V 80mA; 6·3V 3A; 4V 2A
L1, 2, 3 See text

Mounting assembly for V9. Panel-mounting unit for F1. 6in × 12in × 2in aluminium chassis with base plate. Ten small pointer knobs. Coaxial, mains and wander plugs. Eight Noval valveholders, seven skirted with screening cans. Wood and material for cabinet. Tag strips, mains lead, wire sleeving, etc., etc.

Component Values for Videoscope Modifications

Altered Values of Components existing in Original Design:
- R19 now 6·8kΩ 1W
- R23 now 6·8kΩ 1W
- R34 now 33kΩ $\frac{1}{2}$W
- R35 now 33kΩ $\frac{1}{2}$W
- R36 now 1kΩ $\frac{1}{2}$W
- R43 now 4·7kΩ $\frac{1}{2}$W
- R69 now 2·7kΩ $\frac{1}{2}$W
- R72 now 2·7kΩ $\frac{1}{2}$W
- R75 now 270kΩ 1W
- VR2 now 2kΩ 1in
- C20 now 0·1μF paper 500V
- C21 now 0·1μF paper 500V
- C25 now 0·01μF paper 500V
- C54 now 0·1μF paper 1kV
- L2 now 575μH
- L3 now 575μH
- V9 now DG7–32/01 (Mullard)

Components in Original Design now discarded:
- R73, R74, R76, C55, C56

Additional Components:
- R92 100kΩ 1W
- R93 4·7kΩ $\frac{1}{2}$W
- R94 10MΩ 1W
- R95 1MΩ $\frac{1}{2}$W
- R96 10kΩ $\frac{1}{2}$W
- R97 1kΩ $\frac{1}{2}$W
- R98 2·2kΩ $\frac{1}{2}$W
- R99 100Ω $\frac{1}{2}$W
- R100 1MΩ $\frac{1}{2}$W
- R101 1·5kΩ $\frac{1}{2}$W
- R102 10kΩ 1W
- R103 4·7kΩ 1W
- R104 33kΩ 2W
- R105 2·2MΩ $\frac{1}{2}$W
- VR9 5kΩ 1in
- V10 ECC81
- C61 3pF ceramic
- C62 3pF ceramic
- C63 100pF ceramic
- C64 100μF electrolytic 30V
- C65 0·15μF paper 500V
- C66 0·15μF paper 500V
- C67 16μF electrolytic 350V
- C68 0·5μF electrolytic 250V
- C69 0·5μF electrolytic 250V
- C70 2μF electrolytic 250V
- C71 2μF electrolytic 250V
- Tr1 2N1613 (General Electric) 2N3705, 2N916, or other silicon *npn* transistor 30V/$\beta \geqslant 30$
- D9, 10 12V zener diodes (Z12, Brush Crystal Co.)
- D11, 12, 13, 14 56V power zener diode (ZL56, Brush Crystal Co.)
- SK12 Coaxial panel socket

List of Components for Probe

- R92, 93, 94 4·7MΩ 10% 1W carbon
- C61, 62, 63 50pF approximately, 500V ceramic, see text
- Tag-strip, coaxial plug, coaxial cable, p.v.c. tape, etc.

Chapter Nineteen

A COMPREHENSIVE WOBBULATOR

THE introduction of semiconductor diodes which vary considerably in capacitance with applied voltage has enabled some interesting and useful circuitry to be developed. Applications such as remote tuning by d.c. control, automatic frequency control and parametric amplification at v.h.f. are practicable. The writer recently acquired such a diode, and used it to develop a wide-range wobbulator for use with television receivers. The device proved so simple to use that it was thought worthwhile to extend the range of the wobbulator to cover f.m. response curves, and it also proved possible to add the usual broadcast i.f. range of 465kc/s. Thus, a general purpose instrument was designed, and is described here.

Capacitance Law

A preliminary investigation of the properties of the diode showed that if a linear relationship between oscilloscope trace and displayed frequency was to be achieved, the oscilloscope timebase generator could not be used to provide the voltage sweep needed, especially with TV and broadcast band ranges. It was therefore necessary to provide a separate source of variable voltage, with the correct waveform. This necessitated a waveform generator with special characteristics; and, as some complication was going to be incurred in any case, it was decided to make no effort to simplify the instrument, but to develop it into a more useful device by the provision of markers, sync output to lock the oscilloscope firmly, variable r.f. output, and a pre-amplifier and waveform mixer combined. It now became possible to arrange a logical layout of leads to and from oscilloscope and receiver—in fact, the wobbulator is also used as a distribution centre. This obviates the tangle of leads so frequently found when wobbulators are in use.

Sections

The instrument thus consists of the following elements—waveform generator, marker generator, r.f. oscillator, with variable reactance circuit, and waveform mixer. Outputs are provided at low impedance by using cathode followers. Fig. 1 shows in a block diagram how these various units are combined into one instrument.

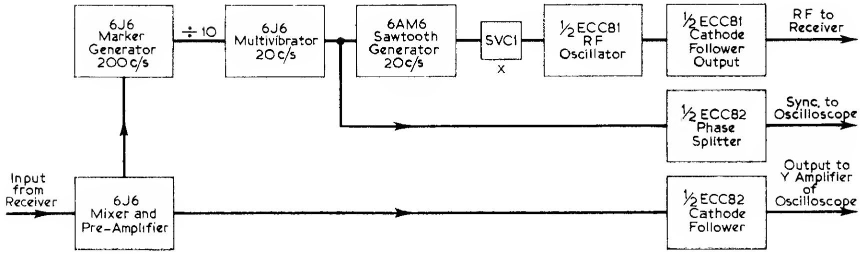


Fig. 1. The block diagram of the wobbulator.

R.F. Oscillator

It will be convenient first to describe the r.f. oscillator and its associated variable reactance circuit—which includes the variable capacitance diode type SVC1 (Mullard)—since the requirements of this stage determine the design of the rest of the apparatus to a large extent. The variable capacitance diode is a semiconductor device in which, when biased in the reverse or non-conducting direction, the capacitance of the "depletion layer", at the interface of *p*-type and *n*-type materials, varies with the applied bias voltage. If this diode is made part of the tuned circuit of an r.f. oscillator, the frequency of oscillation will vary if the applied bias potential across the diode is made to vary. It might be thought at first sight that if part of the X-deflecting voltage of an oscilloscope were applied across the diode a suitable frequency sweep would result. Roughly this is true, but if no precautions are taken certain undesirable features appear. These are listed below.

(a) The frequency sweep is non-linear, being cramped at one end and extended at the other. Thus the oscilloscope trace is misleading, although, if corrections are applied, it can be useful. This effect will be discussed shortly.

(b) Sufficient sweep may not be obtained to cover the required range.

(c) The shape of the response curve may be affected by the speed of sweep, especially where selective circuits occur in the receiver under experiment.

The effect first mentioned above depends on the fact that in an LC tuned circuit, frequency of oscillation depends on the square root of the capacitance, the other parameters being constant.

$$f=\frac{1}{2\pi\sqrt{(LC)}} \text{ or } f=\frac{k}{\sqrt{(C)}}$$

The linearity or otherwise of the relationship between applied voltage and oscillator frequency depends on the relationship between applied voltage V and diode capacitance C. If C were inversely proportional to V^2 the above equation could be rewritten

$$f=\frac{k}{\sqrt{(C)}}=\frac{k'}{\sqrt{(1/V^2)}}=k'V$$

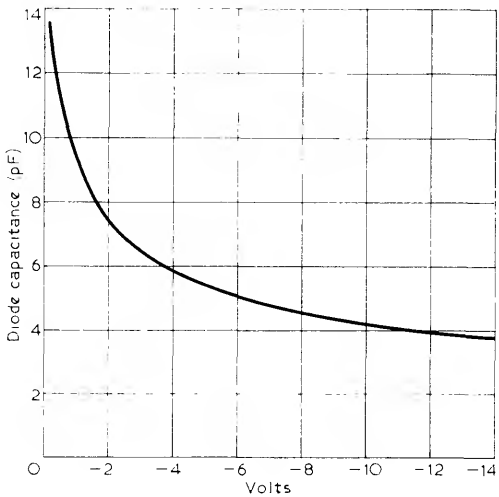


Fig. 2a. The variation of capacitance with voltage applied to the SVC1.

and linearity of frequency sweep would be obtained by applying a linear sawtooth voltage across the diode. The relationship $C \propto 1/V^2$ is approximately true for point-contact diodes, but is not precise enough for junction diodes. Fig. 2a shows the variation of capacitance of the writer's specimen of the SVC1, and it will be seen that there is a major departure from the law $f = k'V$ when, as in Fig. 2b, frequency is plotted against diode voltage.

Consideration of Fig. 2b shows that as the diode voltage drops from a certain value, the frequency drops more rapidly than it should. This effect could be reduced by adding capacitance to the circuit and so swamping the irregularity, but this would minimise the sweep available, by reducing the percentage capacitance change.

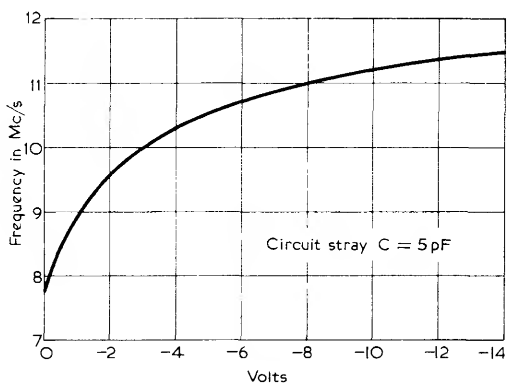


Fig. 2b. Variation of frequency with voltage applied to the SVC1.

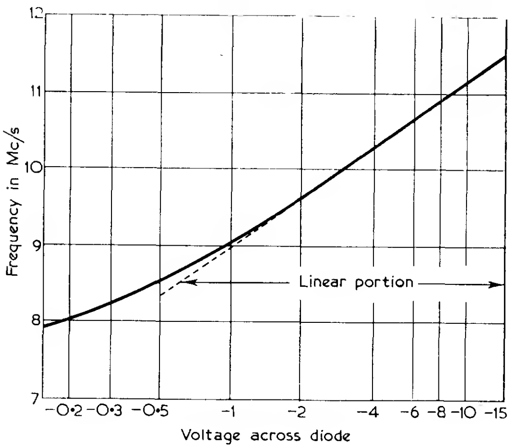


Fig. 3. The SVC1 in a variable circuit—stray capacitance amounted to 5pF.

Exponential Sweep

It will be seen also that if it could be arranged for the voltage to drop less rapidly as it approaches zero, a much more linear frequency sweep could be obtained. A little calculation shows that an exponential drop in voltage gives a great improvement, and the resulting sweep—shown in Fig. 3 by plotting frequency against voltage on logarithmic graph-paper—is practically good enough. A little extra capacitance in the circuit can be afforded, since the sweep is ample, and occurs in any case because of "strays". This also improves results, and by the correct choice of voltage waveform an essentially linear sweep is achieved. However, only one or two pF can be allowed, and thus the aim must be to cut circuit strays to the minimum.

The choice of sweep generator is thus limited to one having ready exponential waveform output. In addition it should be negative-going; in this way, by combining it with a d.c. potential across the diode, the best sweep of frequency can be obtained. The choice is thus for a screen-coupled phantastron, whose source is a low d.c. potential. The usual Miller run-down is very linear when the h.t. line, the source of voltage, is at a fairly high potential. At low potentials, however, the run-down is *slightly* exponential in form, and by adjusting the charging potential to the correct value an excellent correcting waveform can be obtained.

Stray Capacitance

Obtaining sufficient sweep is essentially a problem reducing to that of minimising circuit capacitances. If this is carried out, very little voltage sweep is needed. With the SVC1 the maximum voltage is −20V, which allows for a maximum d.c. potential of −10V with a peak-to-peak sweep of 20V. It proves sufficient in this design to use a d.c. potential of −8V and a

sweep of 16V to cover the required wavebands. It is probable, however, that in the 34–38Mc/s band, the usual i.f. for television receivers, the sound channel cannot be covered at the same time as the whole of the vision band. This is normally quite unimportant.

Sweep Speed

Highly selective circuits cannot deal with fast sweeps, and if the effort is made to use fast sweeps, it is found that the shape of the response curve varies with sweep speed. Ideally, something under 10c/s is preferable, especially where steep-skirted response curves are involved. This, however, requires a.f. couplings, in the various amplifiers, having time-constants of the order of seconds if distortion is to be avoided. In this design, a sweep of about 20c/s is arranged; this is a compromise, certainly, but on the whole not a bad one.

Circuit capacitances are normally reduced to the minimum by using as the r.f. oscillator, a Colpitts circuit. The Colpitts circuit is inconvenient, however, when combined with the reactance diode, and an alternative circuit, the "Cathode Hartley" is here employed. By tapping the cathode into the tuned circuit as near "earth" as possible, the heater-cathode capacitance is reduced in effect. In addition, a relatively small coupling capacitor, connecting the tuned circuit to the valve grid, helps to "stand-off" the tuned circuit from Cag and Cgk.

Marker Circuit (see Fig. 4)

Markers for the oscilloscope trace are generated by a Schmitt trigger circuit—arranged to be free-running. With this circuit, sharp and narrow pulses are obtained which, on differentiation, afford positive and negative-going spikes of negligible width. This multivibrator runs at 200c/s, and is used to count-down to another similar multivibrator running at 20c/s. The output from the latter triggers a screen-coupled phanastron which also runs at 20c/s and provides the sweep voltage across the reactance diode. The 20c/s Schmitt valve also provides a sync pulse, brought out to a terminal on the instrument, to enable the oscilloscope to be locked to the same frequency. By interposing a cathode follower in this lead, the effect of external connections is reduced to negligible proportions, and by providing the cathode follower with an anode load resistor, a choice of either positive or negative sync pulses—to suit the oscilloscope—is available by way of a single-pole, two-way switch.

Mixers

The waveform from the receiver being tested is mixed with the marker pulses by using two valves (each half of a 6J6) with a common anode load resistor. A small amount of pre-amplification is provided by this stage, as this has been found to be convenient when a wide-band Y-amplifier is used

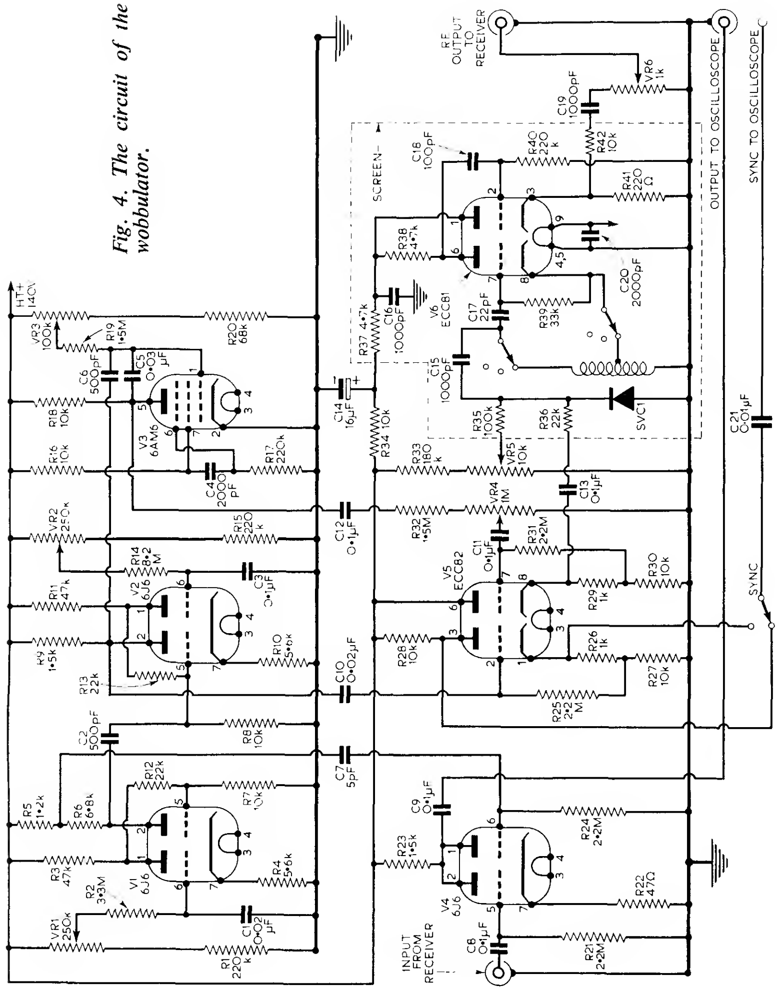


Fig. 4. The circuit of the wobbulator.

in the oscilloscope. The displayed trace is thus divided by ten marker pulses, and it must be stressed that these are *equally spaced irrespective of the linearity of frequency sweep.* In use, the space between the markers is calibrated by using an external signal generator. The method will be described later.

The d.c. voltage across the reactance diode is applied by means of a potentiometer across the h.t. supply—an external control is provided. This is not used to set a critical control point—this is unnecessary since the trace is substantially linear with regard to frequency. It is used in practice to shift the trace to the desired point relative to the markers, which are fixed in position. Thus, accurate alignment with a desired marker can be carried out, and this is convenient where actual measurements are to be taken. Neither is the d.c. control used to tune the oscillator; it could be, but it is best to do any trimming on the actual oscillator coils themselves, leaving the d.c. control in a central position.

The extent of sweep is controlled by varying the amplitude of the sawtooth voltage across the SVC1 diode. For checking work only, any convenient sawtooth voltage can be used, but for measurement the maximum setting should always be used.

Fig. 4 shows the complete circuit diagram. This is by no means as complicated as it may seem, and the layout is far from critical.

The power supplies for the unit, not included with the main circuit (Fig. 4), are given in Fig. 5. They are conventional in the supply of h.t., and 6·3V a.c. at about 3A. The h.t. value of 140V may seem somewhat low, but the reason for this is that if the supply is stabilised by a neon lamp there is plenty in hand. It is essential to remove all hum from the h.t. supply—the aim should be to have not more than 0·1V ripple—and in the construction of the unit care must be taken to keep heater supplies well away from high-impedance grid points. For this reason, grid impedances are kept as low as practicable throughout. Here and there values of 2·2MΩ have to be employed however, and these are the points at which to avoid 50c/s pick-up. Smoothing by a 200μF + 200μF condenser is employed in the prototype, with a resistance

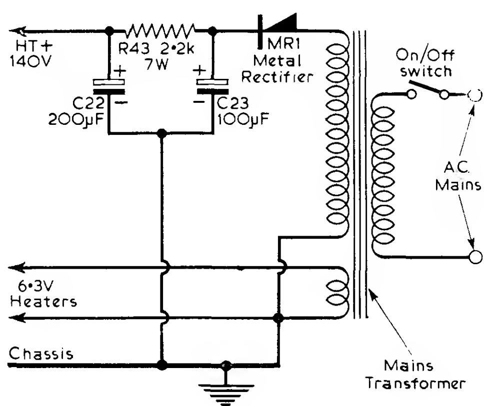


Fig. 5. Circuit of a suitable power supply.

of 4·7kΩ in a 250V a.c. supply to the rectifier. A metal rectifier produces least heat in use and is recommended.

The Schmitt trigger circuits are reasonably stable with regard to frequency, provided stabilised h.t. is used. Multivibrators are inherently sensitive to h.t. variation, however, and if the supply is not stabilised it will be necessary to bring out the 200c/s circuit speed-control to the front of the panel. There is the advantage in this in any case that the number of markers may be varied at will, from about 5 to as many as 20.

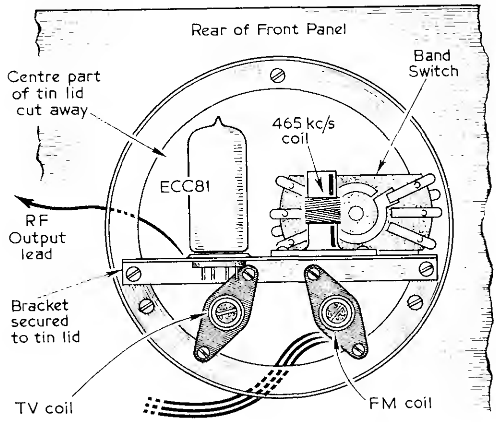
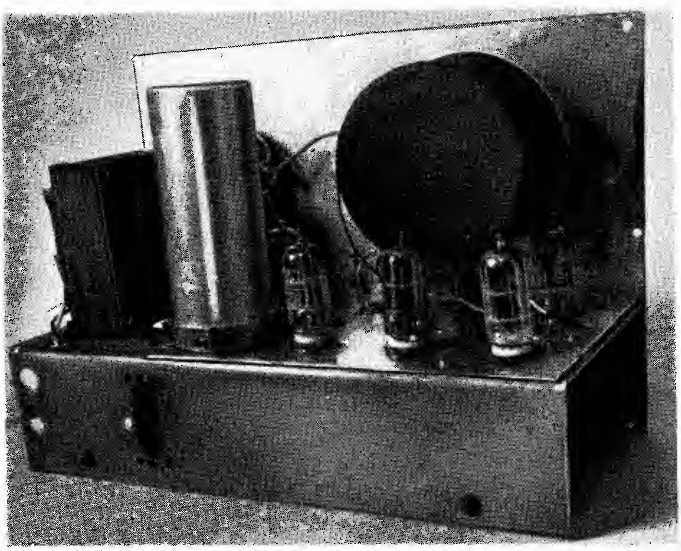


Fig. 6 (above). *The layout of the components inside the screening-can of the oscillator section.* (Below) *A rear view of the wobbulator without the cabinet, showing the oscillator screen.*



Screening

The oscillator section is very carefully screened. Unless the output from any kind of signal generator is to be transferred to the receiver in a rough-and-ready fashion, radiation or induction field must only be allowed to escape from the case at one point, namely the coaxial output point. Otherwise an unknown, and usually large, amount of field escapes and may find its way into the receiver at almost any point. This is no way to conduct experiments!

Correspondingly, thorough decoupling of the oscillator within its screen is needed. Both h.t. and l.t. leads are decoupled, together with the leads carrying d.c. and sweep voltages to the SVC1 diode.

In the prototype, the oscillator screening can consist of an empty "tin" of aluminium 3½in in diameter and 2½in deep. A small chassis is constructed to fit into this, and on the chassis is built the complete oscillator (see Fig. 6). The lid of the tin is fixed to the front panel of the instrument, the centre portion of the tin having been cut out with a sharp penknife, and the chassis is mounted at right angles to it in such a position that, when everything is in place, the tin can be screwed into the lid. Leads are brought out between the lid and the panel—the aluminium is sufficiently ductile to permit this—and

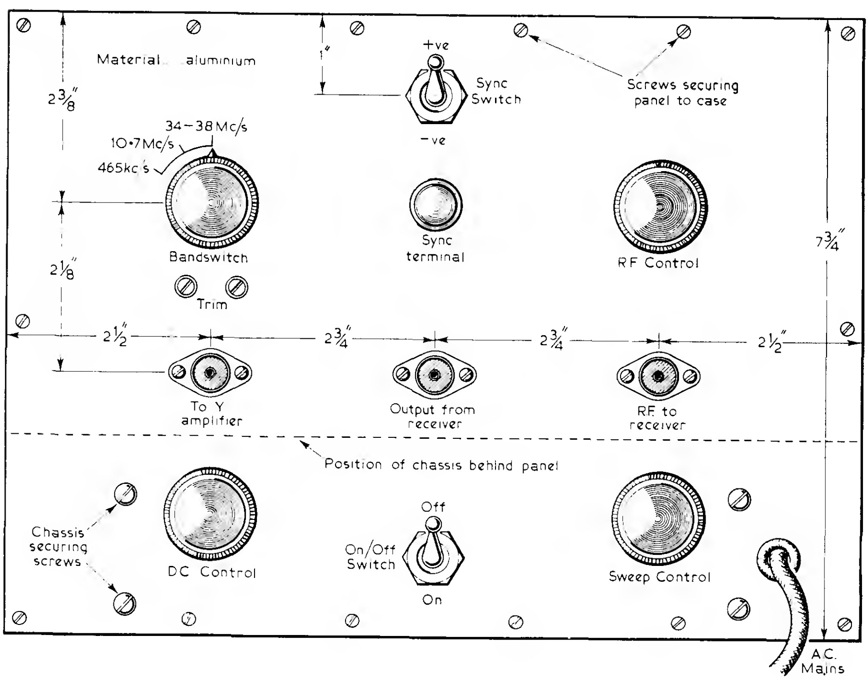


Fig. 7. Drilling details of the front panel.

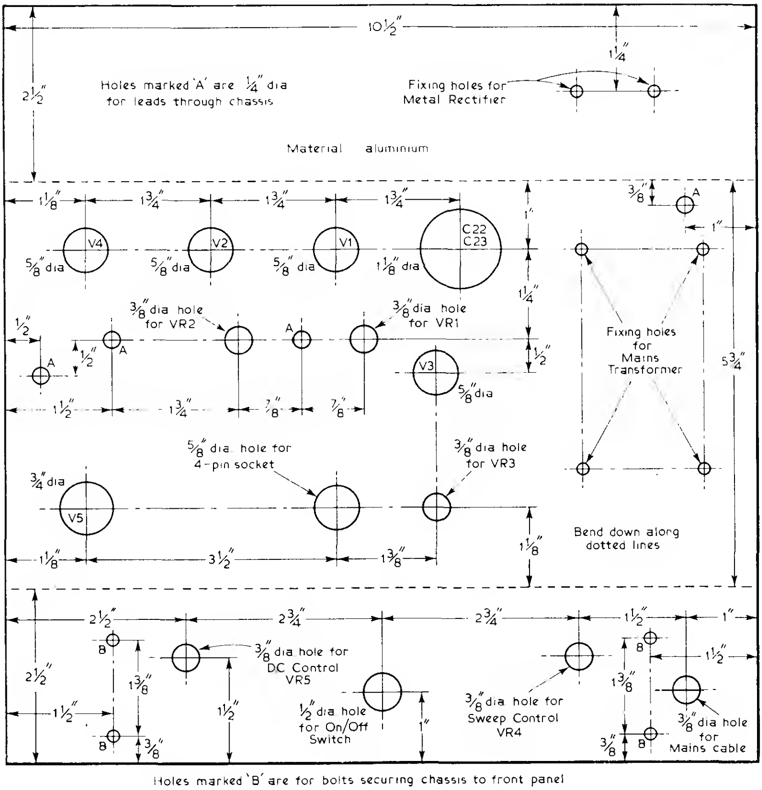
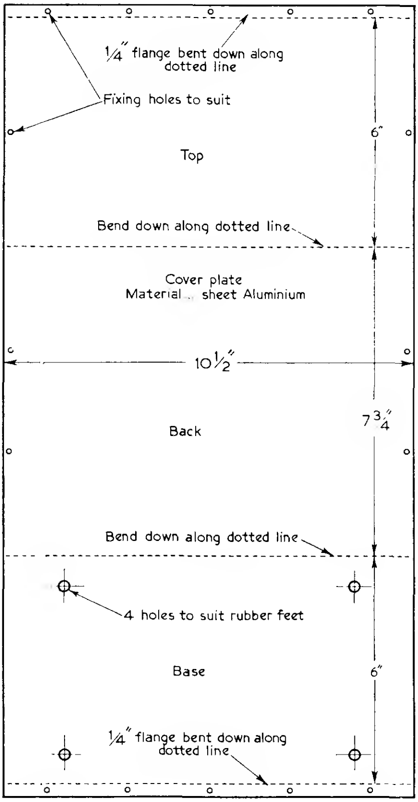


Fig. 8. Drilling details of the chassis.

when the tin is screwed into place, the whole is firmly held and effectively screened. Slots are cut in the tin for ventilation; ideally they would be covered with wire gauze, but this has proved unnecessary since the instrument is built into a complete metal container and thus a double screen is afforded which gives very good reduction of external field. In the prototype, Fig. 7, the mains lead are also decoupled by means of 100-turn chokes and 2000pF capacitors; the only leakage of r.f. field is from the centre spigot of the coaxial connector and the $\frac{1}{4}$in trimming holes which are necessary in the front panel.

The former is by far the larger source of leakage except when a metal screwdriver is used to trim the coil cores—the use of an insulated tool is therefore recommended. The 465kc/s coil is aligned before the can is screwed over the oscillator assembly, and no external trimming hole is needed.

Chassis details are shown in Figs. 7–10, and wiring and switching details in Figs. 11–14.



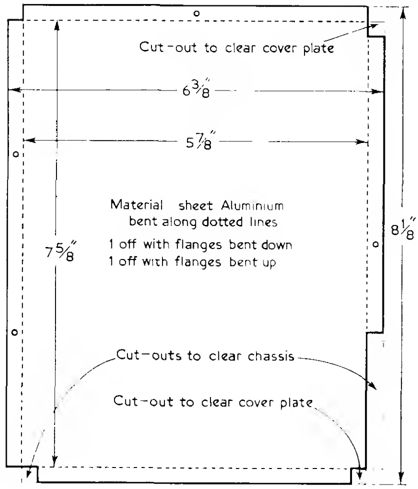


Fig. 9 (left). *Details of the wrap-round case of the wobbulator.*

Fig. 10 (above). *The dimensions of the side-panels.*

Control of Output

It is important for the r.f. output control—or 1000Ω variable resistor—to have zero resistance at its minimum. If it does not, r.f. output will not reduce to a sufficiently low value at the minimum setting of the control. The earth connection must also be of very low impedance. Several wires in parallel, or thick copper braid, must be used and the length of this connection must be kept to the minimum. A wire-wound potentiometer is quite unsuitable for VR6 as its inductance would be appreciable.

A similar low value of minimum resistance must be obtained in the sweep control, or it will be impossible to achieve an unmodulated r.f. output. It is difficult to find a commercial compotent with a low enough resistance and the writer found that an ex-government wire-wound component proved most successful, giving at its lowest setting only 5kc/s sweep. It should be noted that the SVC1 diode, at moderate d.c. potentials, is relatively very sensitive to a small a.c. input.

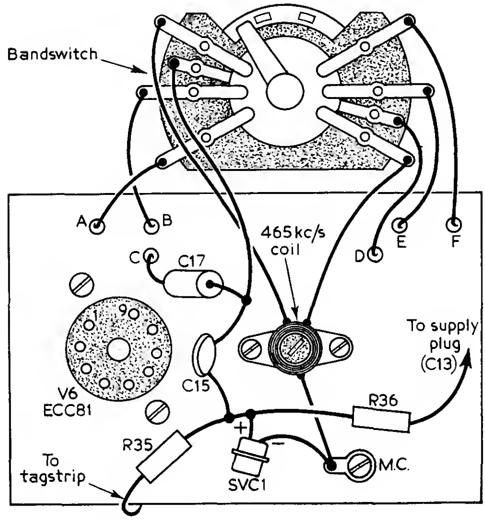
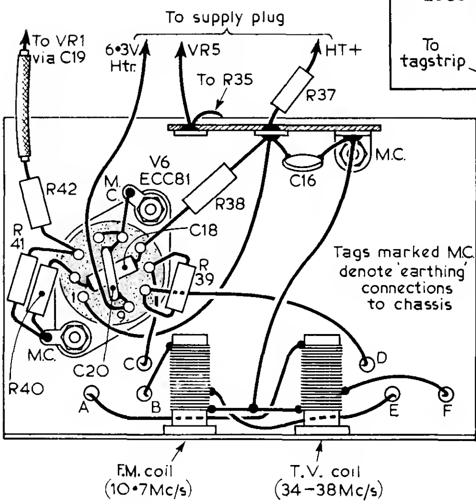


Fig. 12 (above). *The switching of the r.f. oscillator coils.*

Fig. 11 (below). *Wiring details of the oscillator circuit.*



Components

Apart from the foregoing simple precaution, there are no special difficulties about the circuit. All resistors are $\frac{1}{4}$W; the oscillator capacitors of 100pF or less are silver-mica; all the 1000pF capacitors are disc-ceramic types, while the remainder may be of tubular or ceramic type. Layout is not critical in any respect, although of course in the oscillator circuit, stray capacitances must be kept to a minimum by careful wiring.

Alignment

1. *Aligning the 200c/s multivibrator*
   (a) Attach headphones via 0·01μF capacitor between R7 and chassis.
   (b) Rotate VR1, until the note is flat on A below middle C and a little

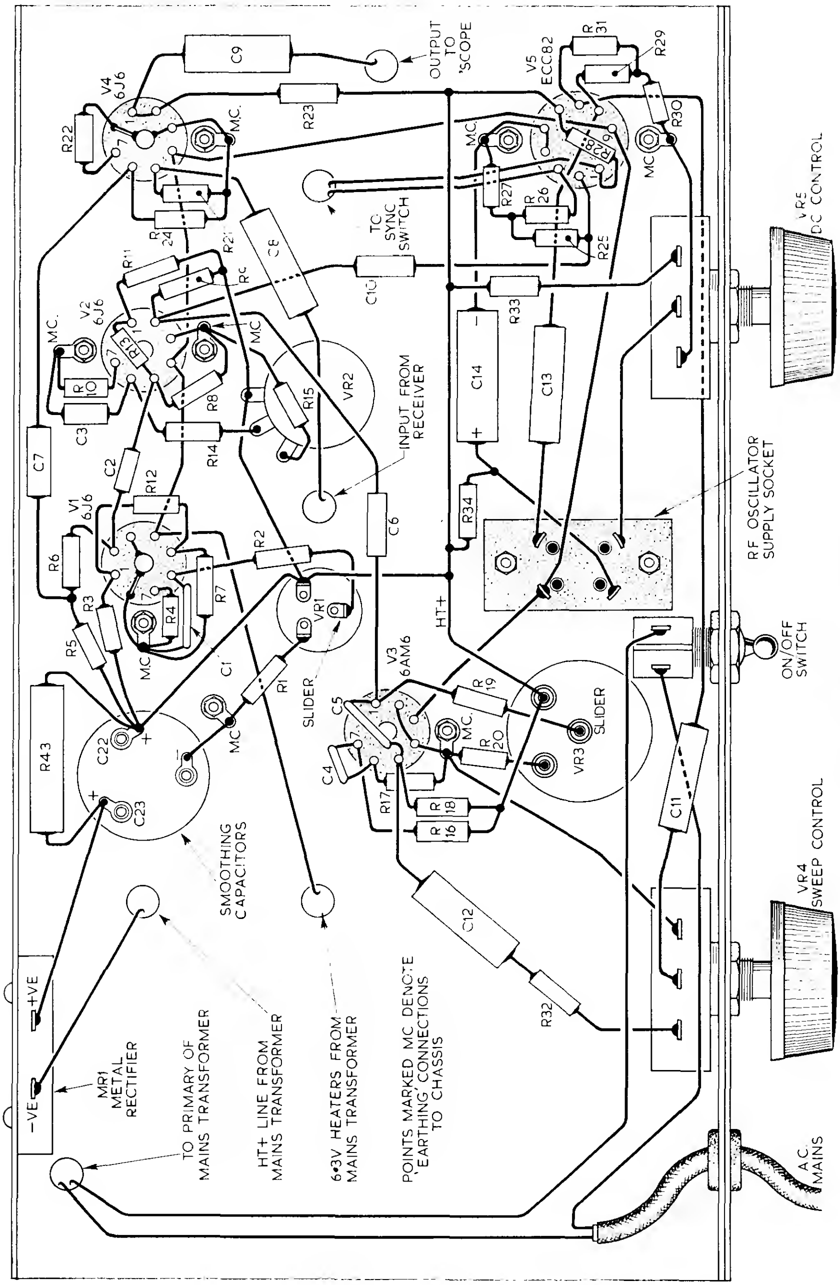


Fig. 13. The underchassis wiring.

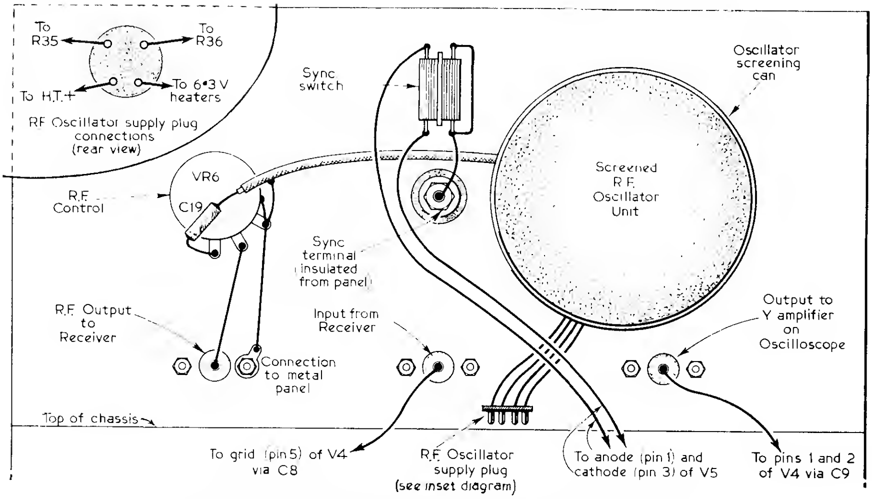


Fig. 14. The connections at the rear of the front panel.
(The inset shows the plug connections for the r.f. oscillator.)

sharp on F below middle C. If an audio generator is available, adjust to 200c/s.

The above gives "near enough" alignment.

2. *Aligning the 20c/s multivibrator and sawtooth generator*
   (a) Attach headphones between R10 and chassis.
   (b) Rotate VR2 until a rapid flutter is heard—not really recognisable as a note.
   (c) Insert 4–6in of bare wire in the "input from receiver" socket, to pick up hum.
   (*d*) Connect Y-amplifier output to oscilloscope, and adjust trace to about 10 sweeps/s.
   (d) An a.c. waveform should now be visible on the scope. Keep it as steady as possible without sync, and adjust to show 4 or 5 cycles.
   (e) Correct the adjustment of the 200c/s multivibrator by rotating VR1 until each a.c. wave shows 4 markers exactly.
   (f) Remove the plug from the "Y-amplifier" socket on the wobbulator and connect the lead instead to the "sync" terminal on wobbulator. Rotate VR2 until over 5 a.c. waves are displayed, 2 pulses are visible and steady.
   (g) Disconnect the lead to the oscilloscope from the sync terminal and attach it to the anode of the 6AM6 (V3). Disconnect the length of pick-up wire from the "output from receiver" socket; rotate VR3 until two perfect sawtooths appear. These should be somewhat "exponential" in shape, negative-going, as shown in Fig. 15. If they are too linear, try connecting a 1MΩ or 500kΩ resistor between the grid of the 6AM6 and its anode, but this need not be tried just yet.

(h) Connect the oscilloscope "external sync" terminal to the wobbulator "sync" terminal, and adjust the oscilloscope trace to 20c/s; use as little sync as possible. The trace should now lock, displaying one sawtooth only.

The sawtooth linearity must now be adjusted and the oscillator aligned before the method of using the wobbulator can be explained.

3. *Adjustment of sawtooth linearity*

(a) Connect the Y-amplifier lead of the 'scope to the cathode of the ECC82 and R29.

(b) Set VR4 at its maximum setting.

(c) Check the shape of the sawtooth wave. It it is still too linear, increase the value of R29 a little, say, to 2·2kΩ.

The linearity should be such that the exponential shape of the wave is just noticeable—a linear sweep is not wanted, but too distorted a wave is not required either. The ideal shape is such that the sloping portion—the sweep—becomes nearly horizontal just at the very end of the sweep. The shape is not very critical, as long as extremes are avoided.

4. *Alignment of oscillator*

(a) Attach 6in of wire to the r.f. output socket to act as a short aerial. Set the r.f. control (VR6) to maximum.

(b) Tune the receiver well away from any station, and turn up the volume control.

(c) Set the "sweep" control (VR4) to about $\frac{1}{3}$ of its range.

(d) Rotate the core of the coil until reception is obtained.

(e) Check that i.f. is now generated by reducing VR4 setting to zero, and test for a heterodyne whistle on all stations.

It may well be that the r.f. output gives too little output, even at maximum, for a well-screened receiver, when stray pick-up is used. If so, connect the r.f. socket direct to the aerial terminal of the receiver via the usual coaxial lead, and adjust the r.f. control for a suitable output. If this fails, connect the coaxial cable direct to the grid of the frequency changer.

The wobbulator is now ready for use.

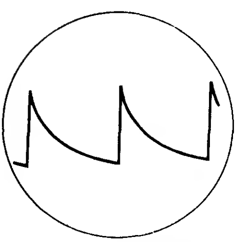


Fig. 15. Sawtooth sweep wave.

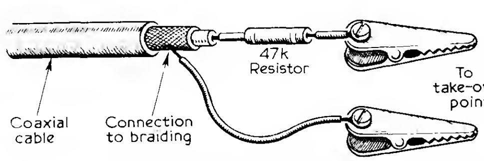


Fig. 16. Terminating the cable for attachment to the take-off point in the receiver.

Method of Use

To avoid any instability, an audio point in the receiver would be best for connecting to the "output from receiver" socket of the wobbulator. To avoid hum pick-up, the signal should be taken from as early a point as possible, best of all from the detector load resistor; this however tends to cause instability by introducing unwanted feedback. To avoid this, a coaxial lead should be terminated by a resistor of value 47kΩ, as shown in Fig. 16. Crocodile clips afford a secure method of attachment; for v.h.f. and TV use, the miniature crocodile clips are best, and the writer has used ladies' hair clips (cut-down) with good effect. Since the detector output is not very great, the usefulness of the built-in pre-amplifier mixer V4 will now be apparent.

If hum pick-up is found to interfere with the displayed response curve, the waveform generators can be made to run at 250c/s (flat on middle C) and 25c/s respectively. Hum is difficult to avoid with mains receivers, but if thoroughly good connections are ensured, and precautions are taken in layout during construction, there should be little interference found. With a TV receiver the video output should be taken from the grid of the video valve, *not from the cathode or grid of the c.r.t.*—the latter is more accessible, but the extra trouble of reaching the earlier take-off point is well worthwhile.

Calibration of Sweep (Fig. 17)

For this process, an external calibrated signal generator is required. The connections are made between wobbulator, oscilloscope and TV receiver, and the signal generator is provided with a short "aerial" at its "high-output" socket. When the response curve is displayed, set VR4 for maximum sweep and centralise the response curve by adjusting the d.c. control VR5. Adjust so that a recognisable point on the curve coincides with a marker.

Marker Pips

Using an unmodulated output, tune the signal generator slowly through the i.f., listening for the sound output (unless the TV vision curve is being displayed!) and watch the response curve closely. A small "pip" will become visible when the receiver begins to respond to the signal generator output (see Fig. 17a).

As the pip approaches the i.f., its amplitude will become larger; when it passes through the peak-response position it may well obscure the response curve altogether, and as it passes out of the i.f. range it will become smaller and finally disappear. It should be clearly visible for several markers. Note the number of markers covering the range of the pip's visibility. From this, the total sweep can be calculated, because the frequency-difference between the pip's appearance and final disappearance is known (from the signal generator).

Fig. 17 (right). *Calibration of the sweep (see text).*

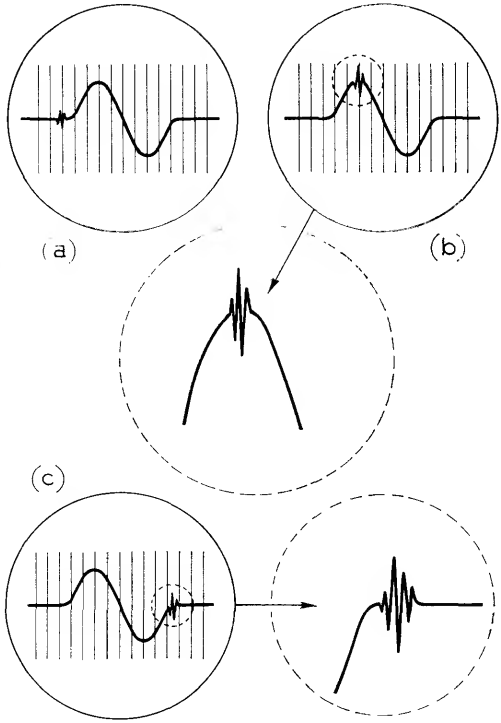
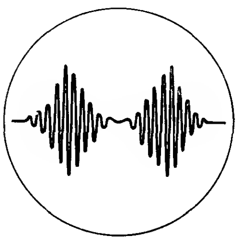


Fig. 18 (below). *If an amplitude-modulated signal is used to produce a marker pip, the pip is of the form shown here.*



A.M. Signal

When an "amplitude-modulated" signal is displayed, at the central point the "pip" will appear similar to that shown in Fig. 18; the "pip" splits into two halves. If the pip blots out the curve altogether, the signal generator output is too high; reduce it by the control provided or by shortening the "aerial".

The calibrations are not the same on all ranges, but once found, they are quite stable and the markers thus have a definite meaning although the meaning is different for each range. The writer's instrument has the following calibrations:

465kc/s—the markers are at 8kc/s intervals.

10·7Mc/s—the markers are at 200kc/s intervals.

34–38Mc/s—the markers are at 500kc/s intervals.

Figures 19–22 show the response of the following receivers:

(A) A v.h.f. transistor portable, with interfering i.f. signal from a short-wave transmitter.

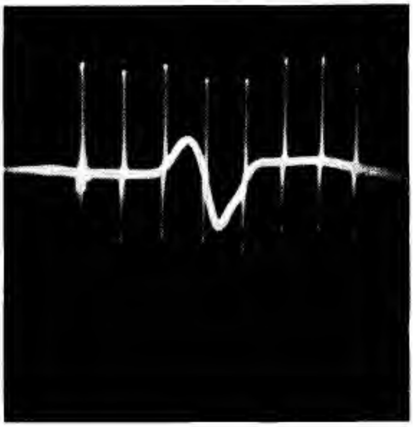


Fig. 19. Discriminator curve for a v.h.f. f.m. receiver (A).

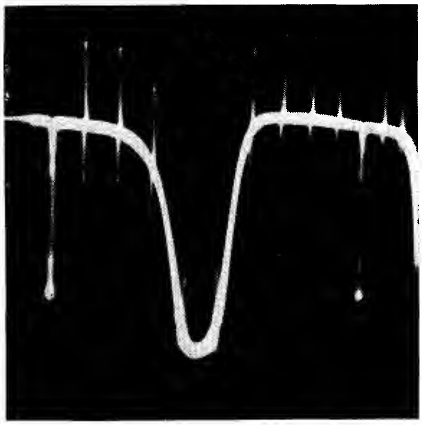


Fig. 20. Response curve for an m.w. transistor portable (B).

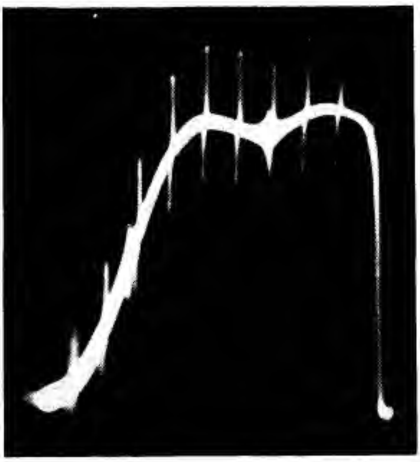


Fig. 21. Response curve of a TV receiver—note the calibration pip (C).

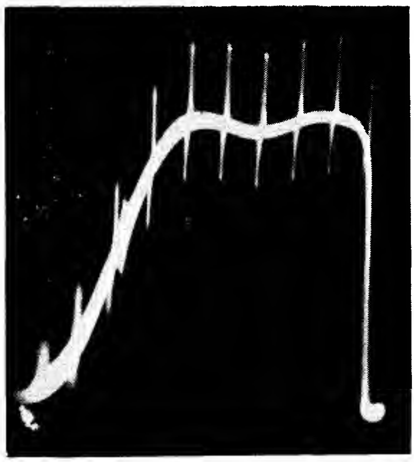


Fig. 22. The same response curve as in Fig. 21 without the calibration pip (D).

(B) Transistor receiver for broadcast wavelengths. The long markers represent the ends of the sweep. This came out the "wrong way up" because of the particular diode connection in the detector stage.

(C) A commercial television receiver—the calibrating pip is in the centre of the response curve.

(D) The same as (C) above without the calibration pip.

In using the wobbulator, care must be taken not to arrange matters so that the d.c. control sets the standing bias on the diode at less than two volts or so. The optimum setting is about 6V or 7V, corresponding to about ¾ maximum, when full sweep is used. The best setting can readily be found by experiment, and the control can then be used to shift the trace a little each way as needed. This is about the only precaution needed in use; the operation of the instrument is very simple and has proved to be virtually trouble-free.

Coil Winding Data

10·7Mc/s 50 turns, tapped at 18 turns from the earthy end, 32 s.w.g. enamelled wire; close-wound.

34–38Mc/s 18 turns of 24 s.w.g. enamelled wire; spaced by the diameter of 32 s.w.g. wire (wires wound together, cemented, and when dry the 32 s.w.g. winding is stripped out). It is tapped at 7 turns from the earthy end.

465kc/s 900 turns of 42 s.w.g. enamelled wire, scramble wound and tapped 200 turns from the earthy end. If this coil is too much to wind, use 400 turns of 38 s.w.g. enamelled wire tapped 150 turns from the earthy end and tune to 1Mc/s or near, so avoiding the I.F. range and injecting signal into the aerial only. This gives equally good results.

All the above coils are on ¼in diameter formers, fitted with iron-dust cores on the 465kc/s and 10·7Mc/s ranges and with an iron-dust core for 34–38Mc/s or brass slug for 40–65Mc/s range, if preferred. Harmonics of this range can be picked up in Band III, and so the overall response curve of a receiver obtained. Naturally, recalibration of the marker pulses will be needed if measurements are to be made.

Chapter Twenty

A C.R.T. TESTER

ONE of the most common problems that the service engineer or amateur encounters is that of the TV set with a low emission cathode-ray tube. The symptoms are that the picture lacks contrast although fully modulated, has low brightness, and the field flyback lines are very noticeable. Increasing the contrast beyond a certain point will turn the picture negative and increasing the brightness causes the picture to go "fuzzy".

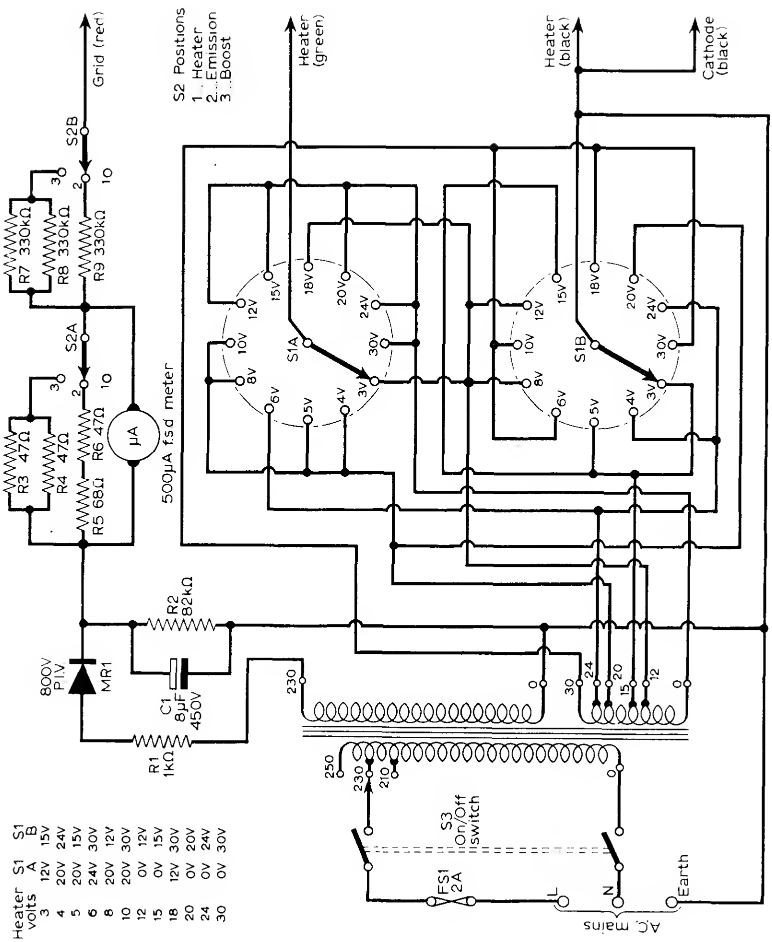
The reason for this lack of "life" is simple—the cathode of the c.r.t. will no longer emit sufficient electrons at its normal operating temperature. This lack of emission can be due to the emitting surface of the cathode having become partially poisoned so that only parts of it are emitting fully. Also, the surface may have almost disappeared, having been worn away by prolonged use. If the first mentioned is what has happened, then there is a good possibility of prolonging the life of the c.r.t. by either permanent or temporary boosting, whilst if the latter has happened a certain amount of extra life may be obtained by permanent boosting.

Function of the Tester-Rejuvenator

The tester to be described checks the emission of a tube when it is suspect, boosts the cathode in an attempt to get rid of the non-emitting portions, and then is used again after boosting to check the new emission. From these operations it is possible to deduce whether temporary boosting was successful or that either permanent heater boosting or tube replacement is desirable. It should, however, be realised that there is no guarantee of success when trying to boost a c.r.t., and there is always the possibility of finishing it off completely. In most cases when a c.r.t. is well past its prime it is worth taking the chance.

Tester Circuitry

It can be seen from the circuit diagram (Fig. 1) that the main component is a multi-tap transformer T1 which is used for supplying the various heater voltages required for c.r.t. testing. These voltages are selected by a multi-way switch, S1A and S1B, and the required voltage is fed to crocodile-clip leads which connect to the c.r.t. heater and cathode. The other lead, which goes to the c.r.t. grid, is either open-circuited or fed from a 300V positive supply via either R7/R8 or R9 (which are in series with a meter).



Heater volts	S1 A	S1 B
3	12V	15V
4	20V	24V
5	20V	15V
6	24V	30V
8	20V	12V
10	20V	30V
12	0V	12V
15	0V	15V
18	12V	30V
20	0V	20V
24	0V	24V
30	0V	30V

Fig. 1. The circuit of the instrument.

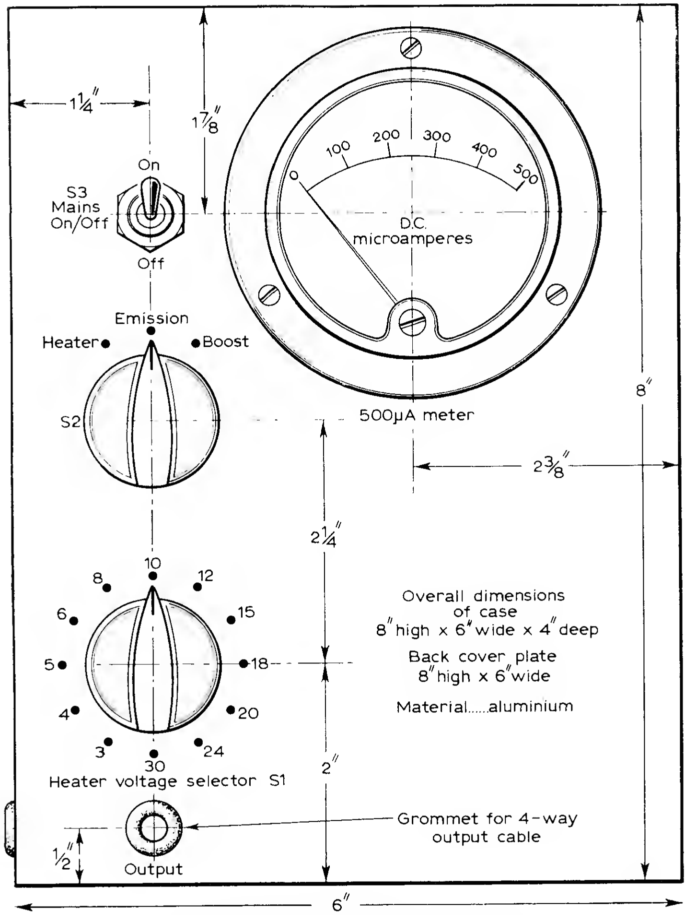


Fig. 2. The layout of the front panel.

Selection of the open-circuit position or appropriate resistor is made by switch S2B, while selection of the meter shunt required in each case is made by S2A.

The 300V supply is derived, through MR1, R1 and C1, from the 250V secondary winding of the transformer. If a suitable transformer is difficult to obtain it should be possible to use two: one with a 250V secondary (or thereabouts) rated at 10mA or more, and one with a low-voltage secondary, tapped as indicated in Fig. 1, and rated at 1A.

Testing for Emission

When checking a suspect c.r.t., the crocodile clip leads are connected to

the appropriate base pins, the correct heater voltage is selected, and the grid switch S2 is put to "Emission".

The grid and cathode of the c.r.t. then act as a diode in series with 330kΩ across a 300V supply. If the cathode is emitting fully, the forward resistance of the "diode" will be low, the current flowing will be approximately 1mA, and the meter will be deflected to full scale. This current is greater than the normal peak white beam current, but a good margin on emission is required if a c.r.t. is to have a useful life.

A low emission c.r.t. is indicated by the grid meter failing to reach anywhere near f.s.d. after running for some time.

Boosting

If from the previous test it is found that a c.r.t. is low in emission, boosting can be attempted.

With the grid switch S2 set to "Heater", step up the heater voltage by one increment of the selector switch for an hour or two. This will cause the cathode temperature to increase appreciably and the emitting surface will undergo a disruption that may cause some of its non-emitting surface to regain its powers of emission.

Check afterwards on "emission" with the normal heater voltage. If some increase in emission is noted, repeat the heater voltage step-up until either good emission is obtained or no further improvement can be had.

Temporary heater boosting is often sufficient to rejuvenate a tube, but if this method is unsuccessful, switch the grid selector to "boost" with the heater voltage stepped up as before. The meter should deflect to full scale after a while. Once this has happened leave the c.r.t. in this condition for ten minutes and then check on "Emission" at the normal heater voltage. This type of boosting strips the top layer of the cathode emitting surface and exposes an unpoisoned layer below if sufficient cathode exists. As before, if some but not enough improvement is found, repeat the treatment until sufficient emission is obtained. Sometimes increasing the heater voltage by more than one increment of the switch can prove effective, but permanent damage to the heater-to-cathode insulation may result.

Construction

As the circuit is simple and the component layout is in no way critical, any chassis or box convenient to the individual builder can be used.

Some guidance for the intending constructor is given in Fig. 2 and Fig. 3 which show the front panel and the interior arrangement.

An insulated crocodile clip is fitted to the end of each of the four output leads and so enables connection to be made easily to the c.r.t. pins.

It can be seen that there is no provision for c.r.t.'s with 2V heaters. This is because in the author's experience c.r.t.s of this type are rarely encountered

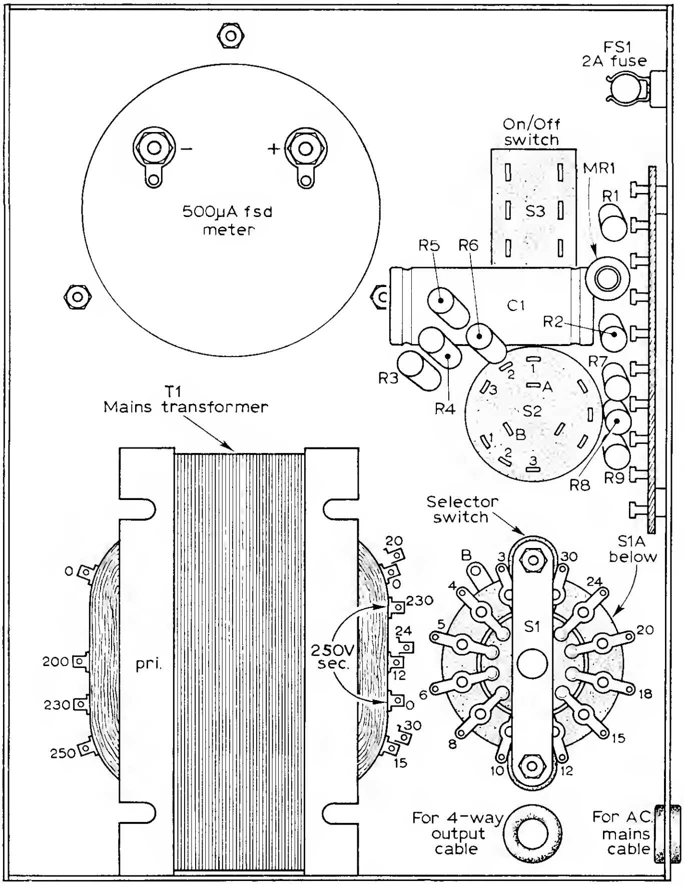


Fig. 3. The layout of the components.

and it was considered preferable to have a higher heater voltage available for boosting emission on line timebase valves and rectifiers.

The author used a 0·5mA f.s.d. meter in the original tester, but any meter of f.s.d. up to 1mA can be used. If the meter resistance is known, then the values of required shunts for the "Boost" and "Test Emission" conditions can be calculated.

With a meter of unknown resistance, the best method to find shunt values is to use a 200Ω variable resistor in parallel with the meter, adjusting this resistor so that the meter just deflects fully when the grid and cathode leads of the tester are shorted together. Fixed resistors equivalent to the values found in this way can then be substituted.

Chapter Twenty-one

A GENERAL PURPOSE Q-METER

IN receiver design, it is always important to ensure that the bandwidth of each tuned circuit is appropriate to its application. This is especially important in receivers intended for the reception of f.m. signals, and, of course, in television receivers, overall bandwidth is very critical. Where coupling between two inductances has to be adjusted to a precise value—a common problem—it is necessary to know the Q of each coil. The experimenter is constantly confronted with the question, "What is the Q of this coil?" and unless he can determine it in some way his design has to be conducted in an empirical manner. This is time-consuming to say the least, and the instrument described here—which was made up very cheaply at the cost of about five hours actual construction time—will be found to pay its way in time saved in a very short period.

The chief causes of energy-loss in coils intended for tuned circuits are dielectric losses in the material included in the distributed self-capacitance of the coil, such as wire insulation and former material, eddy currents in nearby metallic objects—including slug cores if used—and core losses if iron dust cores are involved. Skin effect in the wire of the coil is also important. The normal "ohmic" resistance is usually negligible, though sometimes it has to be taken into account. It will thus be seen that in all practical cases it is hopeless to try to calculate the losses of a coil, even if with special arrangements calculation is sometimes possible. Even more is this the case when the coil is connected to a valve, where other effects occur also to increase circuit losses; with transistors similar effects—but more marked—occur.

Coil losses are usually expressed in terms of an equivalent series resistance damping an ideal coil of no losses. The "goodness" of the coil can then be expressed in the following way. If the inductance of the coil is L and its resistance R, consider a current I flowing through it at resonance. This current develops a voltage $V=IR$, but it is clear that the voltage developed across the inductance must be $V'=2\pi fLI$ and if $2\pi fL$ is greater than R, as is usually the case, the circuit has exhibited the property of "circuit magnification". The ratio

$$\frac{V'}{V}$$

which is thus the circuit magnification is seen to be equal to

$$\frac{2\pi fL}{R}$$

or, expressed simply, as

$$\frac{\text{reactance}}{\text{resistance}}$$

This is the quantity which is with certain reservations known as Q; these reservations include the fact that the self capacitance of the coil has been neglected or, if thought about, it has been decided to charge all the losses in the circuit to the account of the coil. Except in special circumstances this is justifiable.

Q is nearly equal to the reciprocal of the power factor of the coil. It may be noted that with capacitors the power factor is usually quoted, although sometimes Q is used.

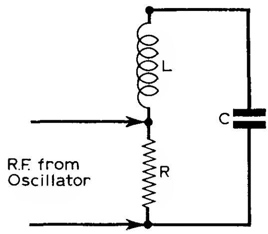


Fig. 1. An explanatory circuit for the Q-meter.

The instrument now to be described uses this effect to measure the Q of a coil. The essential circuit is shown in Fig. 1. L represents the coil under test, C a good quality capacitor the losses of which can be neglected, and R is a small resistor of which the value is known accurately. An oscillator connected across R causes an r.f. current to flow through the resistor and this causes a small voltage to be developed across it. The resistor R forms part of the series circuit LCR, and thus the r.f. voltage is injected into the circuit. Resonance is achieved by adjusting the value of C or by varying the frequency of the oscillator, or both, and a much larger voltage then developed across L is measured by means of a valve voltmeter of very high input resistance.

The r.f. current input has of course to be known and it is very convenient to measure this with a thermocouple milliammeter which can be calibrated accurately in a simple manner. The introduction of the resistance R naturally increases the resistance of the circuit, and so R has to be kept as small as practicable. In commercial instruments the standard value of R is $0{\cdot}04\Omega$; this is not easy for the amateur to manage, and necessitates the use of a heavy current. In addition there is a much worse trouble to be encountered: the reactance of the resistance wire and optional leads. It will be realised that any wire, however short and thick, possesses some inductance. If the wire carries r.f. current a voltage equal to $2\pi f$LI is set up across it, and when it is mentioned that half an inch of 22 s.w.g. wire has an inductance of about $0{\cdot}01\mu$H, it will be realised that the voltage set up by even a small current may not be negligible. If it is not negligible in comparison with the voltage

set up by the resistive property of the wire, errors of large magnitude will occur—not only because of the large spurious voltage itself but also because this voltage is not necessarily in or out of phase with the required voltage.

An example will make this clear. If the resistance is to be 0·1Ω, then the reactance must be such that at the frequency required its value does not exceed 0·01Ω. Suppose the frequency is 50Mc/s, then

$$X_L = 2\pi f L \text{ or } L = \frac{X_L}{2\pi f} = \frac{0{\cdot}01 \times 10^6}{6{\cdot}3 \times 50 \times 10^6} \text{microhenries}$$
$$= 0{\cdot}000032 \text{ microhenries.}$$

This represents a quite unrealisable figure which even in very elaborate instruments cannot be attained. In fact, the Q value at such frequencies has to be obtained by other methods.

Accordingly, a much larger value of resistance is used in this instrument, namely 1Ω. This resistor is made up from a measured length of resistance wire, is wound non-inductively, and is mounted on a support of special construction of which the inductance is about 0·001μH. The instrument can then be used with good accuracy to a frequency of 10Mc/s.

However, the introduction of as much as 1Ω into a tuned circuit is sufficient to render the observed value of circuit magnification much less than Q, unless Q is quite low. Fortunately, the means needed to apply the appropriate correction also affords an alternative means of measuring Q at higher frequencies. This is in fact a calibrated variable capacitor.

At the risk of repeating the well known, it will be as well to have a look at the very simple mathematics of the circuit. Let r be the intrinsic r.f. resistance of a coil of which the inductance is L, and R the inserted standard resistance. The circuit magnification observed is A, while C is the total capacitance of the circuit (including the capacitance of the calibrated variable capacitor at its setting) and $\omega = 2\pi f$.

Then
$$Q = \frac{\omega L}{r} \text{ and } A = \frac{\omega L}{r+R}$$

So
$$\frac{Q}{A} = \frac{r+R}{r} \text{ and } Q = \frac{A(r+R)}{r} \text{ or } A\left[1+\frac{R}{r}\right] \qquad (1)$$

Now
$$Q = \frac{\omega L}{r} \text{ so } r = \frac{\omega L}{Q}$$

Substituting in equation (1) above
$$Q = A\left[1+\frac{QR}{\omega L}\right] = A + \left(\frac{ARQ}{\omega L}\right)$$

Cross-multiplying,
$$\omega LQ = +ARQ$$
$$\omega LQ - ARQ = A\omega L$$

or
$$Q = \frac{A\omega L}{\omega L - AR} \qquad (2)$$

From this equation it can be seen that if the circuit magnification A, the operating frequency, the inductance of the coil and the added standard resistance are known, Q can be calculated simply enough.

It is quite usual for the inductance of a coil, when wound, to be known only very approximately. Hence equation (2) is best rearranged so that capacitance is involved rather than inductance; it is usual for a single variable capacitor to do duty for tuning over all ranges of frequency, and its calibration is not a difficult matter, though of course it needs some care and time. The rearrangement is carried out by remembering that at resonance

$$X_C = X_L \text{ or } \omega L = \frac{1}{\omega L}$$

Substituting in equation (2)

$$Q = \frac{A}{1 - AR\omega C} \qquad (3)$$

From equation (2) or (3) it will be noted that if the added resistance R is very small $Q = A$. This is the way it would be most desirable to use the instrument, but it is out of the question, unless Q is relatively low. However, the formula enables the true value of Q to be obtained from the observed quantity A. The formula should always be applied when A is found to be 60 or more. Fig. 2 is a nomograph from which the value of $AR\omega C$ can be found easily and with sufficient accuracy. The use of this will avoid the tedious calculations of $AR\omega C$, and the chance of getting the decimal point in the wrong place.

The circuit diagram of the Q-meter is shown in Fig. 3. It consists of a small power oscillator in which the range required is selected by means of a wafer switch. Tuning is accomplished in the usual way with an air-spaced capacitor of good quality. In the prototype, this component had a maximum capacitance of about 80pF, but a larger one would not be unsuitable except perhaps on the highest frequency range, 30–50Mc/s. The oscillator has provision for amplitude stabilisation. This is necessary to minimise the amount of harmonic current generated, and the stage operates without grid current. If harmonics are present in any great quantity, it will be seen that the reading of the r.f. milliammeter will include them, but the tuned circuit picks out the fundamental—the frequency to which it is tuned—and so reads a lower voltage than is accountable to the r.f. current flowing. The arrangement used works very well, and has the advantage of being simple. It has however the property of employing negative feedback, and because of this it is possible that at the lower frequencies audio oscillations may be set up which modulate the oscillator. This is no disadvantage in this particular application.

The ranges to be provided are at the discretion of the constructor. Details of the coils used in the prototype will be given, and the ranges used may well be found to be generally suitable. If not, similar proportions may be used in other coils wound to the constructor's own needs. There is little point in attempting to go to much higher frequencies as accuracy is not very likely to be obtainable.

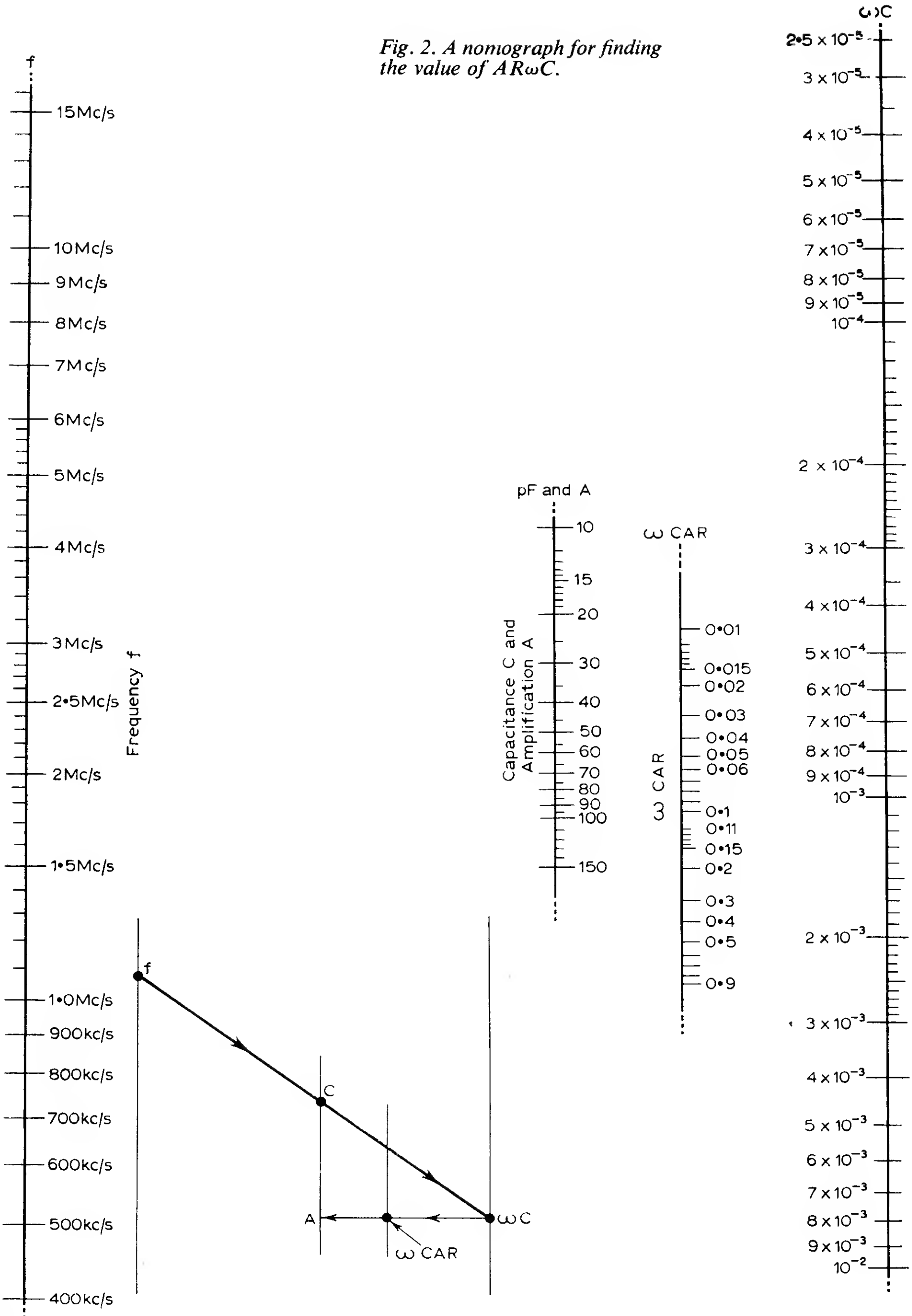


Fig. 2. A nomograph for finding the value of $AR\omega C$.

The valve voltmeter consists of a vacuum diode with a very high input resistance and its output is amplified by a triode arranged in a cathode follower circuit. Drift of this d.c. amplifier is exceedingly small, and zero stability is very good also. The number of ranges of voltage provided may occasion some surprise. The reason for this provision is that the valve voltmeter has many uses, and can be used independently of the Q-meter for any purpose desired. The actual ranges provided will depend on what indicating meter is available. The prototype utilised a 0–25μA movement obtained cheaply as government surplus. This enables a low range of 100mV to be added; it will not in general be used for the operation of the Q-meter, but is very useful for other purposes. If an instrument of 500μA f.s.d. is available it will be found to suffice.

The instrument used for measuring r.f. is worthy of some comment. Small thermocouple meters are available very cheaply, but they seldom if ever read as low a current as is needed in this instrument. Generally they are intended to have a full-scale deflection of about 0·5A to 1A. The current required here is however only about 30mA to 50mA. To measure this small current, a vacuum thermocouple is needed, and such devices usually have a maximum output of 15mV at about 3mA. Thus, what is needed to read properly with a vacuum thermocouple is not a sensitive movement but one with a low internal resistance. In practice, the thermocouple may be removed from a "surplus" instrument and the latter used with the vacuum thermocouple instead. Most amateurs will possess a thermocouple type of meter which has been burned out; this is ideal for the job, as long as the movement itself has not been damaged. The thermocouple specified for the Q-metre was obtained from L. Glaser & Company of 1–3 Berry Street, London EC1.

The colour coding of the leads from the thermocouple used in the Q-meter must be noted carefully. The green coded leads are for r.f., while the red and black leads are for the meter movement. The heater resistance is 10Ω, the couple resistance 2Ω, the couple output 15mV open-circuited, while the maximum current is 50mA. With thermocouples, the overload capacity is very limited—here only 50%. Fuses cannot be used to protect them, and therefore very great care has to be exercised. The measured capacitance between the heater and the thermocouple is only 0·4pF, and hence in the low impedance circuit used with it, leakage of r.f. into the meter movement is not a problem.

The above discussions will have indicated that in this instrument the constructor will have not only a Q-meter but also an accurate valve voltmeter, a test oscillator, a resonance indicator, a means of measuring the inductance of coils, a calibrated variable capacitor and incidentally a meter for measuring low r.f. currents. This seems good enough value for the small cost of building it!

Notes on Construction

The layout of the various circuits (Figs. 3 and 4) comprising the Q-meter is by no means critical, providing a few reasonable precautions are taken.

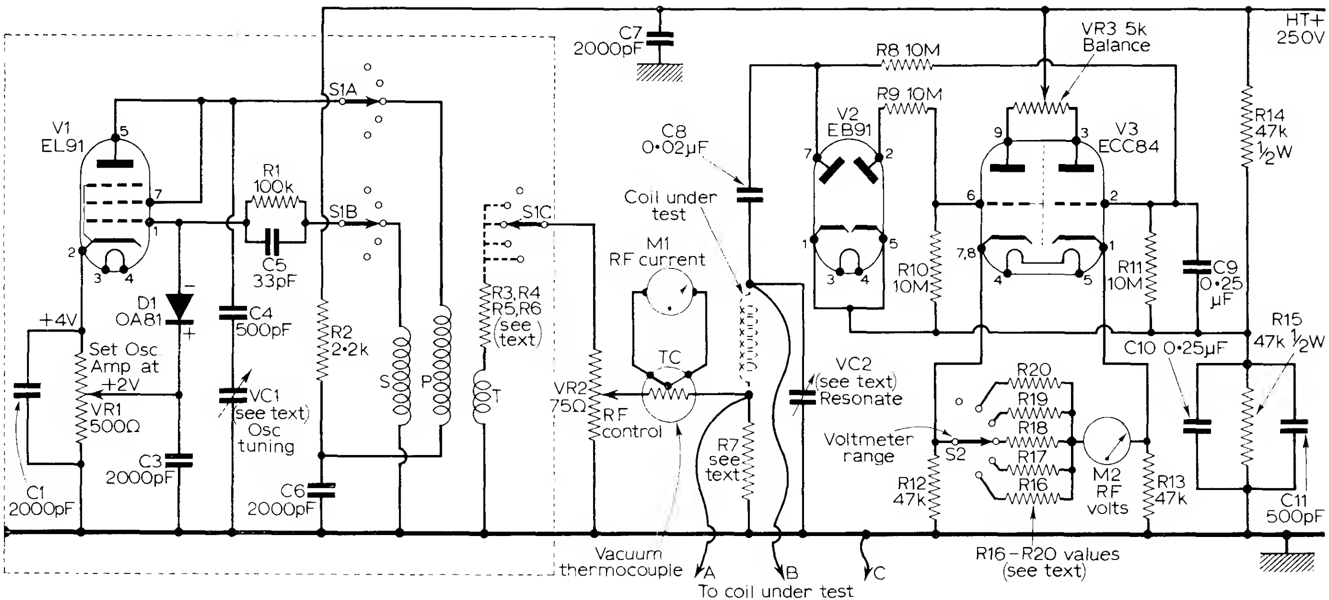


Fig. 3. The circuit of the instrument. For List of Components see page 287.

The valve voltmeter, being purely "d.c.", can be put anywhere so long as it is arranged that its input terminal is close to the capacitor which resonates the coil under test. The oscillator must be very well shielded, and in the prototype occupies a totally enclosed box at one end of the chassis. Very good decoupling is also needed to prevent r.f. from leaking into the remainder of the circuit, whence it can be radiated or fed inductively into the coil under test. There is not quite the same need for perfect shielding that is encountered in a signal generator, so it is here quite allowable for metallic control spindles to emerge from the oscillator compartment, and double screening is not needed.

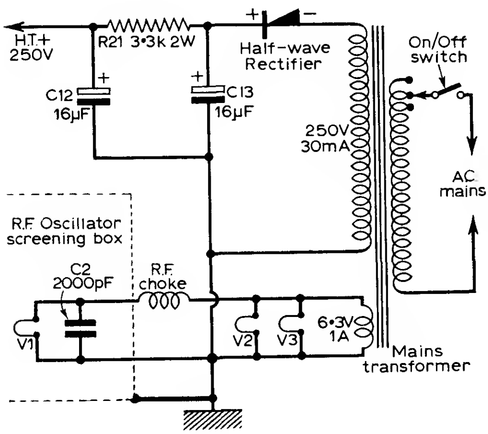


Fig. 4. The circuit of a power supply suitable for the instrument. For List of Components see page 288.

Printed-circuit construction has been used in the prototype, and a suitable circuit template is shown in Fig. 5. Those who do not wish to use this type of construction will however find that a metal chassis serves quite well instead.

If printed-circuit construction is employed, it will be found necessary to mount the valves on the same side of the board as the main components if the use of very long spindles is to be avoided and the depth of the instrument retained within reasonable limits. To accomplish this economically, ordinary valveholders were stripped of the metal surround, and the socket tags were soldered direct on to the connections on the circuit board. This involves remembering, when wiring up, that the order of the valve pins is apparently mirror-imaged. Inspection of Fig. 5 will show what is meant.

The oscillator valve has to supply a maximum of about one hundred milliwatts of r.f. power to the thermocouple. If used in a conventional grid-current oscillator, a small r.f. pentode of the EF80 class would do well, but here the oscillator has to work as a class A amplifier—which is much less efficient. Consequently a small output pentode is used—an EL91, which is obtainable quite easily and cheaply. This is worked well within its rating, strapped as a triode, and so overheating is minimised and with it the frequency stability is improved.

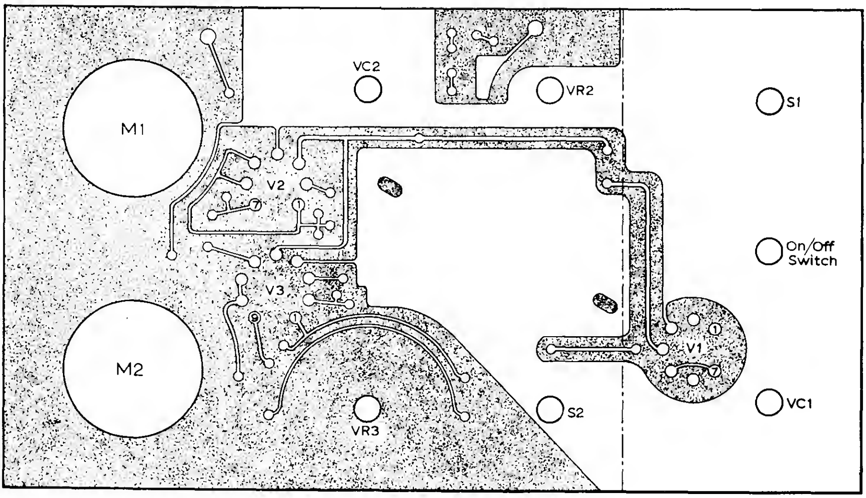


Fig. 5. The printed circuit.

The operation of the amplitude-limiting arrangement is as follows. Referring to Fig. 3, the normal bias due to VR1 is 4V which biases the valve as a Class A amplifier. The slider of VR1 is set at about 2V. As oscillations build up, nothing happens until the oscillation amplitude exceeds 2V, when diode D1 begins to conduct. The current through the diode passes also through the resistor R1 and develops a negative voltage across it which is added to the bias. This moves the working point of the valve towards a region of lower mutual conductance, and the amplitude of oscillations therefore begins to decrease. A steady state is eventually reached and, if the slider of VR1 has been properly adjusted, the valve acts as a Class A amplifier, working without grid current. In fact this does not quite hold good, because diode current flows instead which clips the peaks of the grid waveform. Some harmonic content of the anode waveform is therefore found, and indeed more than might be expected because the "triode" characteristic curves are not quite straight lines in any case. However, the resulting anode current is of sufficiently pure "sine-wave" form for reasonable accuracy. Harmonics can certainly be found, but they are only small and may need a quite sensitive receiver to locate.

The utmost care is needed in constructing the r.f. current measuring section, and especially in making the standard resistor and its mount. The lead between the vacuum thermocouple and the standard resistor should be short—not more than about half an inch, for preference; this is not to avoid capacitive effects, which can be neglected in this part of the instrument, but to avoid the lead's being an appreciable fraction of a wavelength. If it were, the current through the resistor would be different from that recorded by

the thermocouple meter. The standard resistor (Fig. 6) is constructed as follows. Two pieces of thin sheet copper or brass each 0·5in square are cut and placed side by side on a flat surface. The distance between them is adjusted to be 61·7mm as exactly as possible, and they are then pinned in position. A piece of 36 s.w.g. Eureka wire is now cut to a length of about three inches, carefully straightened with the fingers, and laid across the copper lugs squarely and about 0·1in below the top end of each sheet of copper. The resistance wire is then soldered to the copper, taking great care that the solder does not run on to the wire between the copper sheets (Fig. 6). If carefully constructed this standard resistor should be accurate to better than 2%. The wire is now coated thinly with shellac varnish or polystyrene cement and allowed to dry well. When quite dry but not brittle, the wire is folded exactly in half, and twisted gently but firmly so that an even and closely wound pair is formed. The loop at the end can then be closed up carefully with pliers. The idea is to ensure that the twisted wires enclose no area. The twisted wire is now gently bent back and forth upon itself to minimise the space occupied.

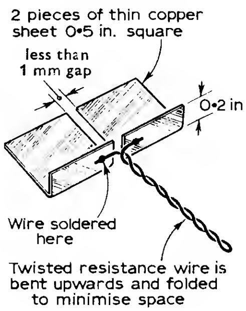


Fig. 6. The construction of the standard resistor.

Each copper lug can now be bent at right angles so that the part to which the resistance wire is soldered is 0·2in high at most. If one of these lugs is soldered to the printed circuit board "earth surface" and the other—soldered to an isolated area of the laminate—is used as the r.f. connection to the thermocouple and the lead to the coil under test, the inductance of the mount can be exceedingly small. The lugs should be so placed as to be parallel and edge to edge, with not more than a millimetre between their edges. If a metal chassis is used, one lug can be attached to the surface by three or four small screws—one is by no means enough—and the other fixed to the chassis by a sufficient layer of contact adhesive or by sticking it to a small paxolin square bolted to the chassis. The lead to the coil under test and that to the thermocouple should be attached to the same point and as near to the point of attachment of the resistance wire as possible. Fig. 7 shows what is required. When in position the thermocouple may be calibrated, using d.c. and a d.c. milliammeter. The greatest care must be exercised to avoid overloading the thermocouple.

It should be noted that Eureka resistance wire is as easy to solder as

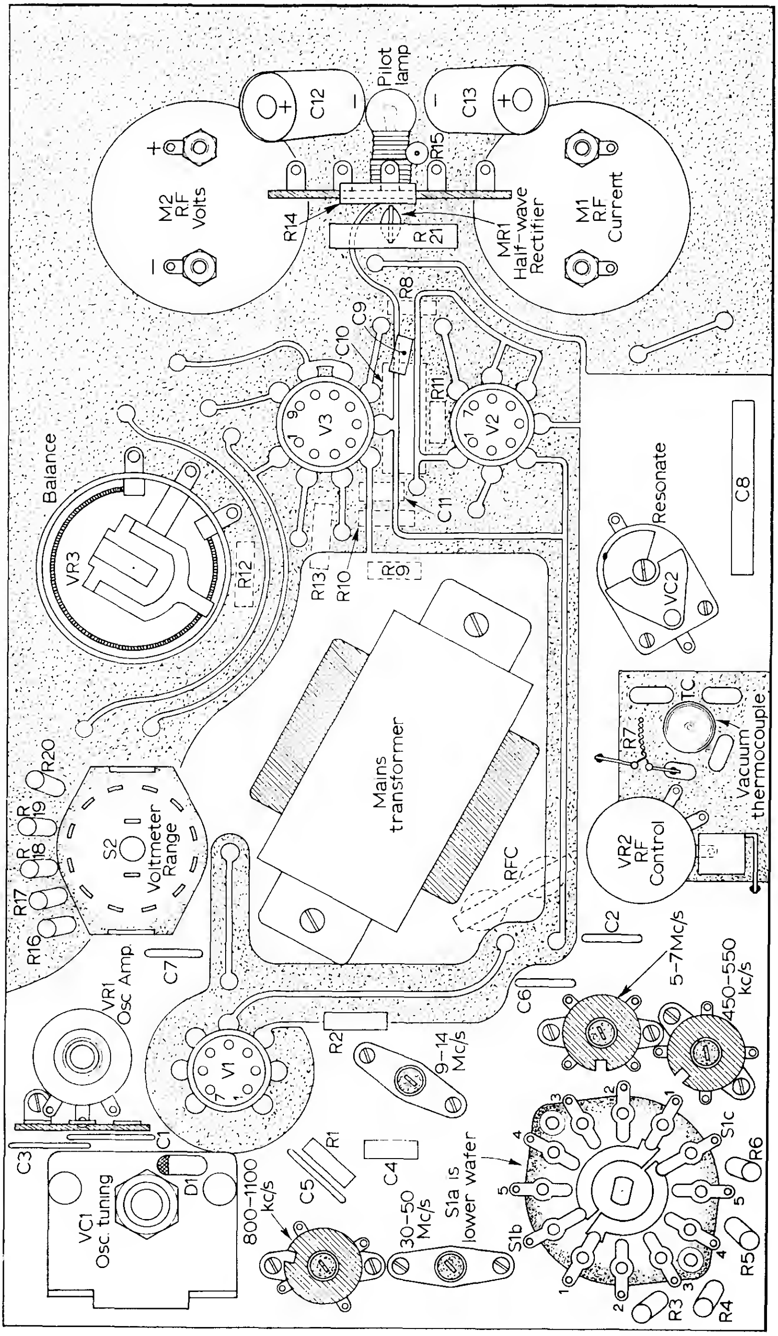


Fig. 7. The complete wiring diagram.

copper. Manganin and any of the nickel-chromium alloys should be avoided as these are hard to solder and need special treatment.

The capacitor connecting the coil under test and VC2 to the valve voltmeter should be absolutely above reproach. If it leaks the slightest amount, a lare voltage may be applied to the grid of one of the triodes in the valve voltmeter and this will prevent any reasonable operation of the meter. A mica or polystyrene film capacitor must be used in this position, as even good paper tubular capacitors show slight sporadic leakage which causes the needle of the voltmeter to waver about instead of remaining in a steady position. Two $0{\cdot}01\mu F$ mica capacitors in parallel will do well in this position.

The heater choke isolating the oscillator from the other valves has to carry a fair current, so may be constructed from 22 s.w.g. enamelled wire. Sixty turns in two pies, each of 30 turns, gives good results. The capacitor C2 is placed just inside the oscillator compartment and the choke just outside.

The values of R16/R20 will depend on the resistance of the meter and its sensitivity. The calculation is simple, and is carried out as follows.

The total resistance in the meter movement circuit is

$$\left(\frac{2}{gm}\right) + Rm + R_{ext}$$

where gm is the working mutual conductance of the triodes, Rm the resistance of the meter and R_{ext} the external resistance in series with the meter. Since, in this instrument, only half the voltage developed at the diode is applied to the grid of the amplifying triode—by means of the potential divider R8/R11—the current through the meter I_m is given by

$$I_m = \frac{0{\cdot}5V_{in}}{2/gm + Rm + R_{ext}}$$

where V_{in} is the voltage input via C8. This, on rearrangement, gives the formula

$$R_{ext} = \frac{0{\cdot}5V_{in}}{I} - \frac{2}{gm} - Rm$$

If I_m is the full scale deflection of the meter, and V in is the maximum reading for the range in question, these can be put into the equation and so the required resistance computed. A value is selected from stock which is within 1%, and the calibration can then be checked using 50c/s supply; the potential divider used for this purpose should have a low resistance—say, 10,000Ω. Suitable ranges are 250V, 100V, 10V, 1V or 2V, and if the meter has high enough sensitivity a range for 0·1V or 0·25V as well. The normal use of the instrument as a Q-meter will involve only ranges 0–10V and 0–1V; the other ranges are provided to allow the valve voltmeter, used separately, to have ranges suitable for other measurements. When measuring r.f. in such circumstances, VC2 should always be set at its lowest capacitance.

The measured input resistance of the meter is 3MΩ, and this puts negligible damping on any practical coil in the ranges of frequency used. Frequencies

lower than 300kc/s, with coils of 1500μH or more, represent a special problem.

The last feature to be described is the means of measuring Q on the highest range of frequency. In this range, the reactance of the mounting and of the standard resistor and of its mounting is too high, and spuriously large values of Q will be obtained by using the equations given on page 275. The capacitor VC2 has to be calibrated in any case if it is intended to apply the correction formula to measured values of Q on the lower frequency ranges. This is accomplished by ensuring that VC2 is of the "straight line capacitance" law, when placing the order for it, and then it can be assumed that if calibration checks are made at 10pF intervals and the law is found to be obeyed approximately, subdivisions of the 10pF intervals on the scale—made by trial and error using a pair of dividers—are as accurate as the 10pF capacitor used for establishing the check points. It is possible to obtain close tolerance capacitors reasonably cheaply, and if two or three of these are used as a check on each other, quite high accuracy is obtainable. The method of calibrating the capacitor VC2 is the obvious "substitution" method; a resonance indicator is available in the valve voltmeter section of the instrument, and the inductor used as part of the tuned circuit should be such as to tune to 465kc/s approximately. At this order of frequency, the inductance of the leads to the coil is so small as to be negligible. It is important also to obtain the residual capacitance of VC2 when set to its minimum position. This can be done in the same way. The scale for the capacitor should be marked in values of total capacitance.

The best results are obtained with VC2 of low maximum value; about 30pF is suitable, and then the 1pF points on the scale are about 6° apart. This enables estimates to be made with fair accuracy to 0·1pF. A suitable capacitor for this position in the circuit is Jackson Brothers air-tune C804 capacitor which is supplied in various capacitances including 30pF.

To measure Q, resonance is obtained at about half-setting (Cx) of VC2. The r.f. control is advanced until the needle of the voltmeter gives a reading of a convenient amount, say 100. Then, the capacitance in the circuit is increased until the meter reads 0·71 of its maximum, and the capacitance (CY) read. The value of VC2 is next decreased until the meter reading, having passed through the peak again, has again dropped to 0·71 of the maximum reading. The reading of VC2 (Cz) is again noted. Q is calculated from the formula

$$Q = \left(\frac{2\mathrm{Cx}}{\mathrm{CY} - \mathrm{Cz}}\right)$$

When obtaining Q in this way it is best to use lead B with a separate earth lead connected to the chassis. This cuts out of circuit the standard resistor R7 and the coil Q is thus read direct without having to make a correction for the added resistance. The calculation above, which can usually be done mentally, assumes that the total capacitance in the circuit is that read from

the calibrated capacitor. This is not strictly true, since all coils have some self capacitance amounting to a few pF. This may be guessed near enough from the dimensions of the coil and added to the total capacitance in the circuit, unless the latter is very small or the coil is of unusual construction. A small solenoid coil will have a self capacitance of about 1pF to 3pF, while a wave-wound coil of higher inductance will usually have a self-capacitance of 3pF to 5pF; it may be larger if the coil is wide and shallow. Normally, it is not necessary to make this correction.

It might be thought that when measuring Q at the highest frequencies without the use of lead A there would be no coupling between the oscillator and the coil under test. If the oscillator were perfectly screened there would still be a quantity of energy transferred since the thermocouple and its leads are necessarily in fairly close proximity to the resonating capacitor VC2. This is enough in the prototype, but if a chassis is used, it may be necessary to ensure that sufficient coupling exists by connecting leads A and B together and using the combined lead with the separate earth lead as the connections to the coil under test. A and B should be connected by means of a capacitor of about 1pF, external to the instrument.

Measurements on the other ranges carried out by the two methods show close agreement. On the 465kc/s range, the Q of a certain coil by current and voltage was found to be 75, while, by the capacitance-variation method, the value was 76. On the medium-wave range, another coil showed values of 88 and 89 respectively. Values on the other two ranges were 63 and 64 (at 5·5Mc/s) and 108 and 103 at 10·7Mc/s. It may therefore be taken that the potential accuracy of the instrument is of the order of a few per cent.

The Q of an aerial can also be determined with this instrument. Generally the actual measurement is less important than the bandwidth of the aerial, which can be derived from Q. Since Q for an aerial is usually very low, neither method described above is very satisfactory, and the following method—though apparently less accurate—may be found sufficient.

If a dipole is under test the coaxial lead attached to leads A and B should be terminated with an 80Ω resistor. This stimulates the actual input of a receiver and also serves to short circuit any hum voltages which would seriously affect the working of the instrument. Advance the r.f. control until, with VC2 at its minimum setting, a resonance point can be found on rotating VC1, the tuning control of the oscillator. Set the valve voltmeter to an appropriate range (1V), and advance the r.f. control until a maximum reading of the valve voltmeter of 100 is shown, at resonance. Note the frequency reading. Swing VC1 each side of the resonance point until a voltage reading of 71 is obtained, and note each reading. Then the bandwidth of the aerial is the difference in the two side frequencies f_1 and f_2, while the Q of the aerial is the resonance frequency f_0 divided by f_1–f_2. For a plain folded dipole without parasitic elements, the terminating resistors should be 300Ω, but a folded dipole with directors and reflector will need 80Ω, as with a plain dipole.

The power supplies for the instrument are modest, the requirement being

for 250V at about 30mA and 6·3V at about 1A. The h.t. smoothing need not be very elaborate. In the prototype, an instrument transformer with a 250V secondary was used, rectification being "half-wave" through a 1000V p.i.v. semiconductor diode. Smoothing was accomplished with a reservoir capacitance of 16μF, a smoothing resistor of 3·3kΩ and a smoothing capacitor of value 16μF (see Fig. 4).

List of Components (Fig. 3)

Resistors:

- R1 100kΩ
- R2 2·2kΩ
- R3, R4, R5, R6 see Table A
- R7 see text
- R8 10MΩ
- R9 10MΩ
- R10 10MΩ
- R11 10MΩ
- R12 47kΩ
- R13 47kΩ
- R14 47kΩ $\frac{1}{2}$W
- R15 47kΩ $\frac{1}{2}$W
- R16, R17, R18, R19, R20 see text

All $\frac{1}{4}$W unless otherwise stated

- VR1 500Ω variable
- VR2 75Ω variable
- VR3 5kΩ variable

Capacitors:

- C1 2000pF
- C3 2000pF
- C4 500pF
- C5 33pF
- C6 2000pF
- C7 2000pF
- C8 0·02μF mica
- C9 0·25μF
- C10 0·25μF
- C11 500pF
- VC1 see text
- VC2 see text

Valves:

- V1 EL91
- V2 EB91
- V3 ECC84

Meters: see text

General notes on the components employed in the instrument can be found in the text

Notes on Capacitors and Resistors:

- C1, C3, C6, C7 Disc-ceramic type
- C5, C12, C13 Mica or silver-mica
- C8 Mica or polystyrene film
- C10, C11 High-quality tubular
- Resistors are all $\frac{1}{4}$W except R14 and R15, which are $\frac{1}{2}$W
- VR1, VR2, VR3 may be wire-wound or carbon-track types
- R8, R9, R10, R11 should be wiped with a clean rag moistened with carbon tetrachloride before soldering into place, and not touched by hand thereafter

Valveholders:

- V1 Any suitable
- V2 and V3 Ceramic or p.t.f.e.

Leads A, B and C:

These leads should be as short as possible, p.v.c. insulated, and terminated in miniature crocodile clips

List of Components (Fig. 4)

R21 3·3kΩ 2W
C2 2000pF ceramic
C12 16μF 350V
C13 16μF 350V

Rectifier: E250C50

Mains on/off switch

Transformer:
Mains primary; 250V 30mA and 6·3V 1A secondaries
R.F. choke

Table A

Values of R3 to R6

Range	*Resistor*
30Mc/s–50Mc/s	zero
9Mc/s–14Mc/s	33Ω
5Mc/s–7Mc/s	68Ω
800Mc/s–1100kc/s	100Ω
450Mc/s–550kc/s	150Ω

Table B

Windings of the Inductors

Range (Mc/s)	*Primary (anode)*		*Secondary (grid)*	*Tertiary (R.F. feed)*	*Wire*
30–50	5	Interleaved with	4	2*	22 s.w.g.
9–14	20		8	3*	28 s.w.g.
5–7	45		18	6*	28 s.w.g.
800–1100 kc/s	180		60	10†	36 s.w.g.
§450–550 kc/s	450		150	25†	40 s.w.g.

* Tertiary is thin connecting wire in plastic sleeving.
† Tertiary is 28 s.w.g.
§ Windings are placed, random wound, between cheeks on the former ½in apart.
All wires are enamelled copper.
All inductors are wound on 0·3in Aladdin formers with dust core.
Windings are placed one on top of the other in the order Primary (anode), Secondary (grid), Tertiary (r.f. feed).
Windings are separated by one layer of Sellotape.
Figures in the above table are complete turns of wire.

Chapter Twenty-two

AN IMPEDANCE BRIDGE FOR TRANSISTOR MEASUREMENTS

THE circuit designer who is interested in high frequencies has had for some time now the choice of vacuum tubes or transistors as his active elements. For certain applications, the vacuum device is by far the more convenient of the two, but in receivers the present tendency is to make more and more use of the solid-state device for reasons which hardly need mentioning. With valves a very extensive body of experience exists together with a well worked-over collection of circuit theory. Most of this has been extremely useful to the transistor circuit engineer, but there are differences in emphasis as well as differences in application with which he has to be familiar.

One of the problems which now has to be faced every day is one which hardly bothered anybody who used vacuum tubes. The vacuum tube at moderately high frequencies imposes little loading on a tuned circuit connected between grid and cathode, and even at Band III frequencies the effect is relatively easily taken into account in the few circuits in a receiver which operate at signal frequency. In an i.f. amplifier the effect cannot be completely neglected, but normally causes few headaches since the input resistance of a valve is usually so much greater than that required in the circuit anyway to obtain the bandwidth required. The use of pentodes, with their extremely high anode impedance, results in negligible resistance in parallel with the anode tuned circuit, and so the h.f. pentode is a device offering few if any snags to the designer. The virtual absence of coupling between grid and anode is also a property which simplifies calculations.

Effects

With the transistor, the effects which could be neglected formerly, now take on the highest significance. Where a grid input-resistance of many thousands of ohms was usual, now a base input-resistance measurable in tens of ohms is found, often associated with a capacitance of anything up to 100pF in parallel. Output resistances of kilohms instead of megohms are common, and feedback capacitances of several pF instead of values a hundredth as large. Circuit design therefore has to regard as of overriding importance such quantities as with valves may be minor difficulties.

The author has, in the past, satisfactorily solved circuit problems for

transistorised equipment. However, one problem encountered was the design of a transistorised television receiver. One especially severe difficulty was the obtaining of the correct bandwidth in the vision i.f. amplifier, and this because the transistors selected for use had to be employed well away from the frequencies for which published data was available. Thus it was necessary to obtain by measurement figures for input admittance, output impedance, feedback and transfer admittance and the corresponding phase angles, before a rational approach to stage design could be attempted. To effect some of these measurements a suitable instrument was made up, and this is now described in the belief that other amateurs may find it useful in their own work.

Originally it was considered that a simple measuring device, such as is illustrated in Fig. 1, might prove sufficient to measure input and output impedances. The considerations were as follows—a signal from a low-impedance source (a constant-voltage generator) is fed into the base of a transistor arranged with its proper d.c. bias. The amplified signal is measured with a valve voltmeter, and the reading noted. Then series resistance is added to the input until the reading of the valve voltmeter is just halved. In this case, half the voltage is dropped across the series resistor and half in the transistor, and thus the added series resistance is equal to the input resistance of the transistor. Poor results were obtained with this arrangement, for the obvious reason that the input capacitance of the transistor causes a phase change which is not paralleled externally. However, at low frequencies consistency of results was observed, and at audio frequencies the method is simple and direct.

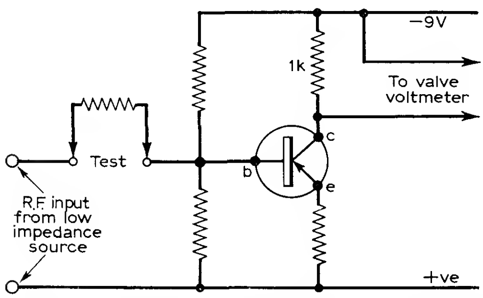


Fig. 1. One method of measuring the input impedance of transistors.

Fig. 2 gives the complete circuit diagram of the instrument as finally made up; Fig. 3 is a simplified version of the circuit which reveals the principle of operation more clearly. Referring to Fig. 3 therefore, it will be seen that in essentials it comprises a bridge. R.F. current from an external source is needed, and this will normally be a well-shielded signal generator with an output of low impedance—usually about 80Ω. When r.f. is fed to the input terminals, it traverses the arms of the bridge R2—R1C1 and R3—R*i*C*i*. R2 and R3 are equal, and so when the impedance R1C1 is exactly equal to R*i*C*i* the potentials at A and B are equal in magnitude and phase. R*i*C*i* is the unknown quantity, and thus if R1 and C1 are known as well as the equality of R2 and R3 the problem is solved. It remains to note only that

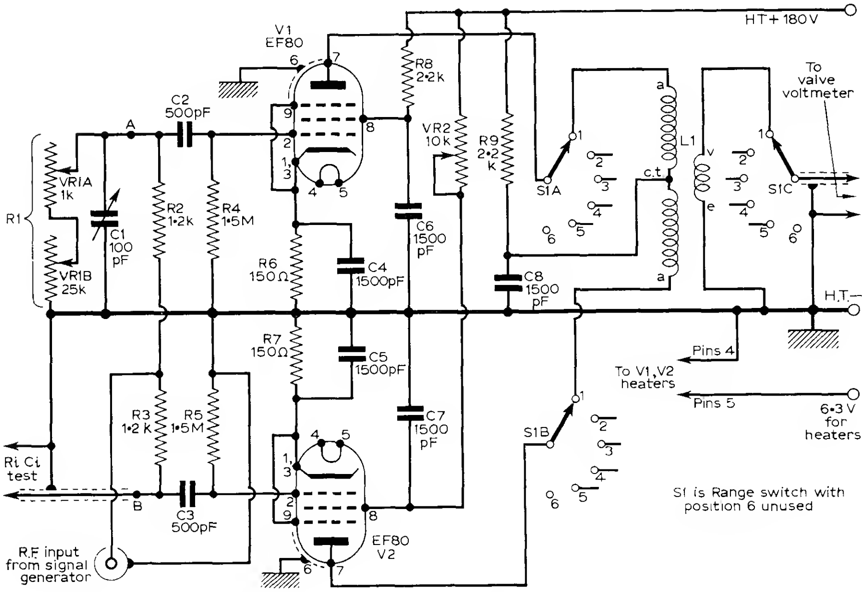


Fig. 2 (above). *The circuit of the instrument.*

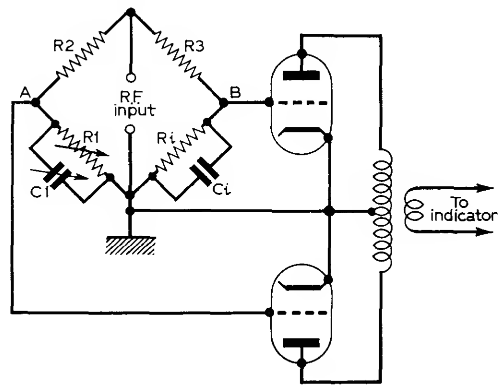


Fig. 3 (right). *A circuit to explain the principle of the instrument.*

the equality of the potentials at A and B is ascertained by feeding these potentials to the grids of two amplifying pentodes, whose anodes supply two separate tuned circuits. To both these tuned circuits is coupled a single r.f. pick-up coil, and the presence of r.f. voltage in the pick-up coil is shown by any suitable indicator. When both pentodes are handling equal signals the fields of the anode coils cancel out and no r.f. is induced in the pick-up coil; when the signals at the two grids are not equal the anode currents differ in phase or magnitude, or both, and a voltage appears in the pick-up coil. Thus by using a reasonably sensitive indicator the precise null point can be reached when R1 and C1 are properly adjusted.

Design Consideration

The basic bridge circuit is of course so well known as to merit little comment. The actual values of components used however may need some explanation. It has to be remembered that the instrument was intended for work with transistors whose input and output resistances are much smaller than those associated with vacuum tubes. A bridge tends to be most accurate when the four arms are nearly equal, and consequently R1, R2 and R3 are rather small. It is for this reason that the r.f. input should preferably be from a low impedance source. On the other hand, the input and output impedances of transistors are so different in size that some compromise is essential if both are to be measured by the same instrument. Thus R2 and R3 are specified as 1·2kΩ, and the balancing resistor R1 is split into two, one of 25kΩ and one of 1000Ω. It is of course necessary to ensure that the 25kΩ potentiometer has a true zero, and a means of ensuring this will be described.

Where a bridge is to be used at high frequencies, it is obviously most necessary to remove as much stray inductance and capacitance as possible. If it cannot be removed it has to be balanced out or, if this should fail, allowed for. While in the instrument described every precaution must be taken to minimise such "strays", certain compensatory devices are used which will be described.

In the first place, leads carrying the measuring current have to be as short as possible. This is not of such import in the anode circuits, but here the danger of stray field affecting the grid circuits exists. Consequently, adequate screening must be provided. Secondly, it is obviously necessary to bring leads outside the instrument to make contact with the transistor or other device under test. Because of this, the leads to R1 and C1 have themselves inductance and capacitance which is not negligible. If the balancing circuit elements R1 and C1 are within the case of the instrument, their leads are likely to have very different values of inductance and capacitance. Thus, when balance is achieved, although resistance will be right, capacitance as measured will show quite considerable errors. To compensate for this difficulty the balancing elements are not connected to the case of the instrument direct but by way of a piece of coaxial cable exactly equal in length to the external lead; the lengths must be correct to 0·1in, including the length of crocodile clips if used.

Besides taking normal precautions such as either matching R2 and R3 exactly, from stock, or using 1% tolerance components, care has to be taken to ensure exactly the same amplification by each pentode. Variable-mu frame-grid valves might be used with an adjustable resistor in one cathode lead. Here however the readily obtainable EF80 is used, and matching of the valves is done by varying the screen voltage of one by a few volts one way or the other. If no precise match can be found it is only necessary to change the valves over and try again; they will seldom be so far from specification that this trick will fail.

Winding the Coils

The tuned circuits in the anodes are very critical, and the most precise symmetry must be achieved in winding the coils. This is not difficult, but requires care and a little skill; it is most pressing where the smallest coils are concerned, for a little inaccuracy in placing one turn cannot usually be corrected in a subsequent turn of wire. It may be found necessary to discard one coil and try again from scratch. The lower frequency coils are very much easier to wind accurately.

Symmetry is further assisted by the actual wiring of the coils into place, and this matter will be discussed later.

Spot Frequencies

Five spot frequencies were decided upon as being all that would be needed for most work. It might have been thought better to tune each of several ranges so as to obtain full coverage of all the interesting bands. However, this would have introduced much complexity—and probably inaccuracy—into the instrument and it was decided against. The actual frequencies used may be decided by the constructor, but it is thought that the following will cover most requirements.

Switch Position—1	10·7Mc/s
2	40Mc/s
3	70Mc/s
4	100Mc/s
5	200Mc/s

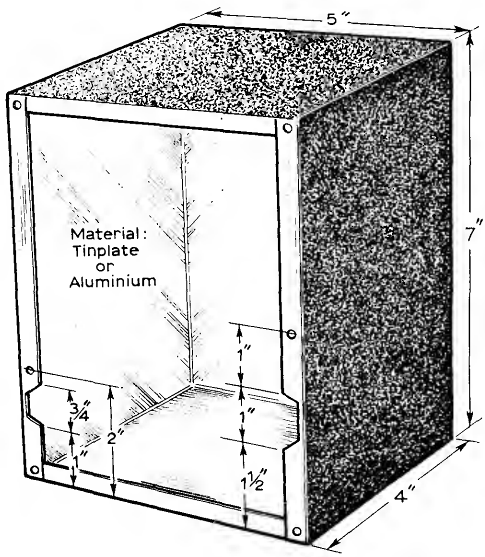


Fig. 4. The construction of the cabinet.

The sixth switch position can be used for an additional frequency or used merely to join the anodes of the EF80's together and to h.t.+ for testing purposes.

Construction

As the circuitry is simple and relatively few components are needed the instrument can be fitted into quite a small box. The chassis was therefore arranged to fit into a small metal container 7in × 5in × 4in deep. The size will of course be determined by the tin box available, but a large container is unnecessary. The lid was discarded in favour of a panel of aluminium, to which the chassis and screening could be attached, so that the whole circuit could be removed at will. To avoid the chance of stray radiation or induction field penetrating the box, the panel was secured by means of several self-tapping screws (see Figs. 4 and 5).

The main wiring (Fig. 6) is carried out on a small chassis mounted at right angles to the front panel, and arranged so that the r.f. input leads are kept very short. This arrangement also ensures that the leads to the transistor under test are as short as feasible. Below the main chassis is a screen which serves to shield the input and balancing components from the anode coils; leads are taken through this screen as necessary by the most direct route. Thus the screens divide the assembly into three; the top compartment contains the valve envelopes and the balancing circuits, the small middle compartment contains the valve wiring (Fig. 7) and the input circuits, while the third contains the anode coils and frequency selector switch together with the valve screen-balancing potentiometer and the terminals carrying the power supplies.

It is essential that the potentiometers which together make up R1 are of the carbon-track type. During the period of development of this instrument it was desired to use a 500Ω potentiometer for one of these, and the most extraordinary results were obtained. Eventually the case was taken off the potentiometer to see what the arrangement was inside; it was a wire-wound type, possessing very considerable inductance. Replacement of this component by a 1000Ω carbon track type removed the cause of the trouble and a very puzzling fault was cleared.

These potentiometers should also be of the smallest physical size that can conveniently be obtained; they have to carry only a minute current and power dissipation is not a problem. Inductance and self-capacitance are serious problems however. Only one at most may have an earthed spindle or slider, and if one has either of these peculiarities it should be placed on the "earth" side—electrically it does not matter which is connected to the chassis, but one must be insulated at least. Neither must be connected to the front panel; earthing is accomplished by means of the "outer" of a piece of coaxial cable equal in length to the coaxial lead to the transistor under test, while the "hot" end of both potentiometers and capacitor C1 is connected to the grid of the corresponding valve by means of the "inner" of

the same piece of coaxial lead. In this way compensation for the length of leads to the transistor is effected. Similarly, the capacitor C1 is not earthed to the front panel but by means of the coaxial outer which also earths the potentiometer R1. Good insulation from the front panel and small stray capacitance to earth are achieved by substantial discs of Perspex or paxolin.

Frequency-selector Switch

For the frequency-selector switch two Yaxley-type wafers are employed, each of configuration two-pole six-way; one of the sets will remain unused.

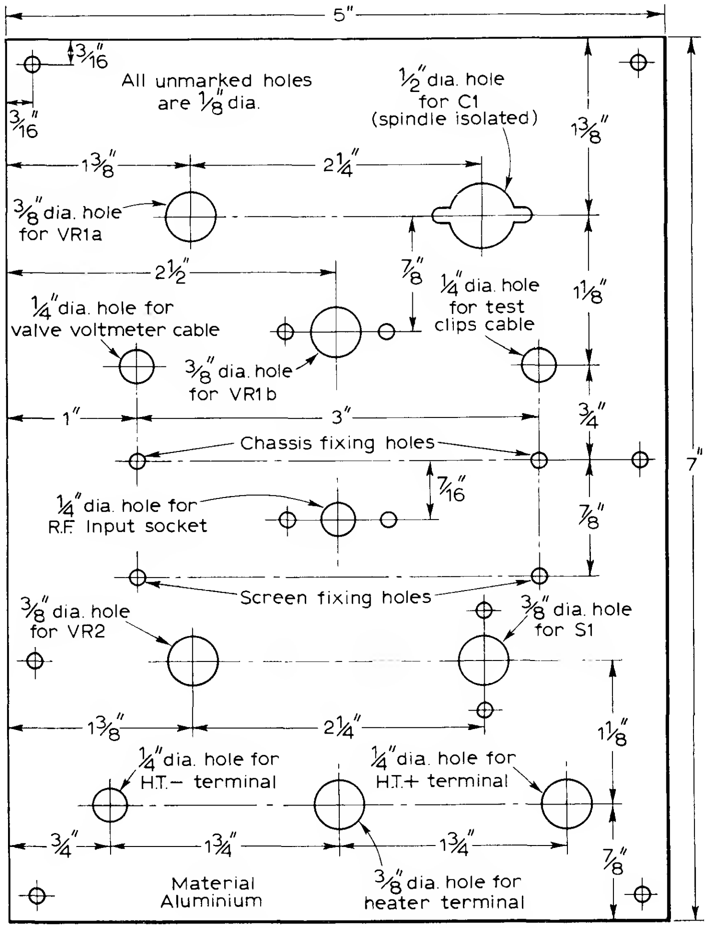


Fig. 5. The drilling details of the front panel.

These are arranged on the selector spindle by means of appropriate spacers to be about 1in to 1½in apart. When wound the anode coils are wired direct between corresponding contacts on the two wafers. The third set of contacts is used to switch the r.f. output cable to the corresponding pick-up coil.

Power Supplies

The instrument has not been provided with its own power supplies, because a suitable supply will generally be available for use with such ancillary equipment not in continual use. However, there is no reason why the constructor should not incorporate a self-contained power pack if desired; it need only supply 0·6A at 6·3V and about 20mA at 180V h.t.

Auxiliary Equipment needed

An external signal generator is needed to supply the r.f. input, and if about 10mV is available the indicator can easily be a valve voltmeter or a crystal diode and microammeter. This combination has the big advantage of not needing an external r.f. receiver tunable to the frequencies used. If it were desired to use the design to measure ohms and fractions of a pF the

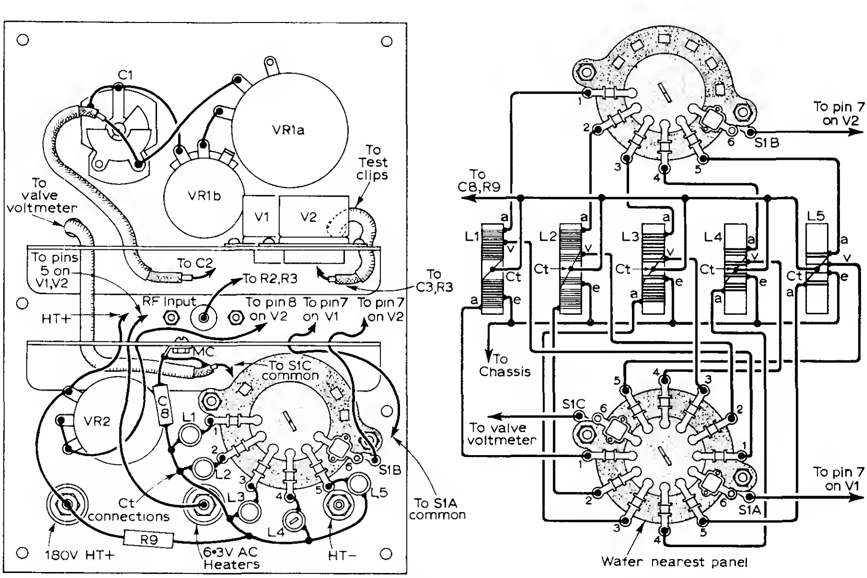


Fig. 6. The main wiring diagram, showing the arrangement of anode and take-off coils.

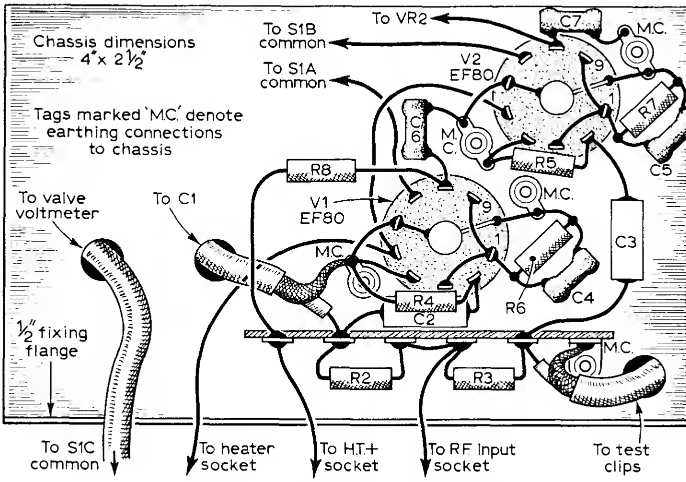


Fig. 7. The wiring of the valveholders.

added complication would be difficult to avoid, but for the purpose intended the bridge is sensitive enough to give good accuracy with the simpler equipment.

Alignment and Calibration

Before inserting into the circuit, both the variable capacitor and the potentiometers should preferably be calibrated against known standards. If this is not practicable, the makers' stated figures may be accepted to a first approximation and—assuming that a "straight-line law" applies—the scales can be divided accordingly.

To set up for use the following simple procedure should be adopted. First, it is required to ensure that both halves of the anode circuit are tuned to the same frequency. To do this, the balancing potentiometers are returned to the position of least resistance and the test leads are left open-circuited. The selector switch is put in the desired position and the r.f. input leads are connected to the signal generator. The output leads are connected to the indicator and power is switched on. After an appropriate warming up period of about ten minutes the signal generator is swung slightly about the nominal frequency until an indication is recorded on the output meter. The signal generator is then tuned as nearly as possible to the peak. A dust core or brass slug is next inserted into one end of the anode coil and screwed in a little, while the signal generator setting is varied a little one side or the other. If the output reading increases the slug is screwed in a little more and the signal generator setting varied, until a position of the slug and the generator can be found which gives the maximum output. If on insertion of

the core the reading first decreases, the core has to be screwed right through the coil until it engages with the opposite winding; again a position is found which gives the maximum output reading. This process is now repeated for each selector switch setting in turn. It may well happen that the tuning point is awkwardly sharp, and if this is found the Q of the tuned circuits should be reduced by connecting a resistor of between 5kΩ and 10kΩ direct between the anode contacts on the wafers. This is unlikely to be found necessary if dust cores are used, but if silver-plated brass slugs are employed it may be useful.

Next, the amplification of the pentodes must be equalised. For this purpose the input terminals are left open-circuited and the coaxial leads to the balancing network R1C1 are temporarily unsoldered. Setting the selector to any desired range an r.f. signal is injected and the potentiometer in the screen lead to V1 is rotated until the output reading is zero or very nearly so. It may happen that to obtain a reasonable zero the pentodes have to be changed over, as mentioned earlier.

The balancing network is now soldered into the circuit. The next step is carried out by short-circuiting the test leads, zeroing the balance potentiometers and the capacitor C1, and then checking once more to see that in this position a null reading is obtained on the output meter. If not, and if it can be assumed that R2 and R3 are equal, it means that one of the potentiometers has not a true zero or else that an unwanted coupling exists between input and output. If the latter, attention must be directed to locating the cause of the trouble. If the former, the balance potentiometers should be removed, the cases opened and a small strip of copper foil soldered to one of the end lugs so that the slider makes good contact with it at the end of its travel. Carbon track potentiometers are especially prone to this absence of a true zero, but the modification needed to correct it is simple to apply.

Taking Readings and Recording Results

To take readings, the method is to zero R1 and C1, leaving the test leads open-circuited. Tune the signal generator to obtain maximum output. Connect the test leads to the device whose impedance is to be measured, and rotate R1 and C1 together until zero or as nearly zero as possible is recorded on the output meter. The resistance and capacitance are then read direct from the calibrated scales.

The result may be recorded immediately as so many ohms in parallel with so many pF, and in this form it is suitable for many purposes. If, however, it is desired to introduce the figures obtained into certain calculations, it may be preferable to convert the readings into admittance, when the relationship

$$Y = G + jB$$

holds where $G = Y \cos \theta = R/(R^2 + X^2)$ and is the conductance while $B = Y \sin \theta = X/(R^2 + X^2)$ and is the susceptance; both are given in ohms. θ is of course the phase angle and X is the reactance of the capacitive component.

Fig. 8a. A view of the front panel fo the bridge.

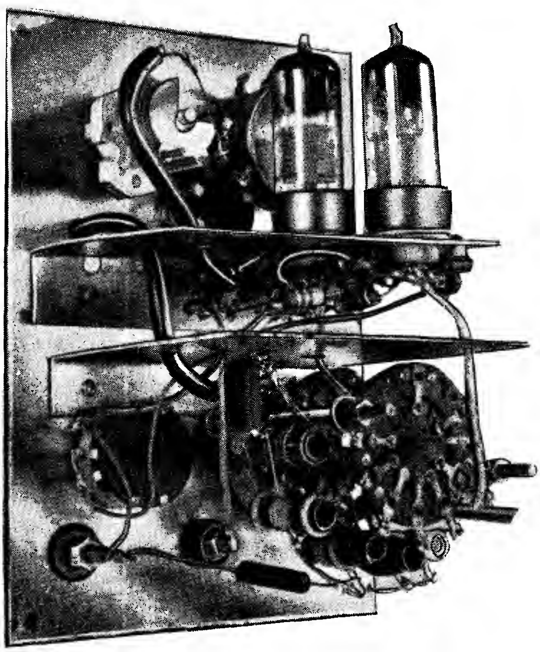


Fig. 8b. A view of the interior of the instrument.

Accuracy

With care in construction and use an accuracy of better than 5% may be obtained on the lower frequency ranges while at 100Mc/s the best obtained has been 7%. The accuracy of 200Mc/s is hard to assess, because the self-capacitance of the resistors R2 and R3 affects matters, while the inductance of test resistors and crocodile clips is not negligible. A conservative estimate might be that with care 15% accuracy may be achieved, and this is probably enough for most purposes.

List of Components

Resistors:
R1 (See VR1a/b)
R2 1·2kΩ
R3 1·2kΩ
R4 1·5MΩ
R5 1·5MΩ
R6 150Ω
R7 150Ω
R8 2·2kΩ
R9 2·2kΩ
VR1a 1kΩ carbon pot.
VR1b 25kΩ carbon pot.
VR2 10kΩ carbon pot.
All resistors ¼W carbon

Capacitors:
C1 100pF miniature air-spaced variable
C2 500pF polystyrene
C3 500pF polystyrene
C4 1500pF ceramic
C5 1500pF ceramic
C6 1500pF ceramic
C7 1500pF ceramic
C8 1500pF ceramic

Switches:
S1 two-wafer type, each wafer, 2-pole, 6-way

Valves:
V1, V2 EF80

Miscellaneous:
R.F. socket. Terminals. Crocodile clips. Two B9A skirted valveholders. Knobs. Materials for panel and box

Coil Winding Data

All coils wound on 0·3in diameter formers with dust cores or brass slugs as needed. The formers should be about 1·25in long. All primaries centre tapped, h.t. to c.t. windings all in same direction.

10·7Mc/s—Primary, 45+45 turns 40 s.w.g. enamelled, spaced 0·3in between ends. Secondary 5+5 turns thin p.v.c. wire close-wound between primary windings. The secondary may have to overlap the inner ends of the primary.

40Mc/s—Primary, 9½+9½ turns 32 s.w.g. enamelled close-wound, spacing about ½in between ends. Secondary, 2+2 turns same wire close-wound adjacent to halves of primary at inner side.

70Mc/s—Primary, 7+7 turns, secondary 1½+1½ turns, as above.

100Mc/s—Primary, 3+3 turns 18 s.w.g. close-wound. Secondary, 1½+1½ turns over-wound at c.t. (thin p.v.c.-covered connecting wire).

200Mc/s—Primary, 1¼+1¼ turns 18 s.w.g. spaced by wire diameter. Secondary, 1 turn as above at c.t., interwound with primary.

Chapter Twenty-three

A TRANSISTORISED VOLTMETER FOR E.H.T.

THE unit to be described was designed to check and facilitate adjustment of television receiver e.h.t. supplies. Fig. 1 shows the circuit diagram, which uses a single transistor as the active element. By this means two main advantages accrue. The first is that the meter and its container are separated from the e.h.t. voltage and the second is that a robust type of meter can be used safely and accurately. In fact, if desired a multimeter can be used instead of an integral meter.

The writer is acutely aware of the need for complete protection of the user from shock and the meter from damage. At first sight it might be thought that a well-insulated high-resistance multiplier attached to an ordinary meter would fill the bill. However, should the meter circuit breaker operate while a test was in progress the whole multiplier would acquire e.h.t. voltage; not only would there be severe sparking across the circuit breaker inside the meter but danger might result to the user. He might get a severe shock in both senses of the word.

Safety is, of course, not the only criterion, although an important one. Accuracy could hardly be achieved by a simple probe-type multiplier because of leakages along the surfaces concerned. These could be avoided by superlatively good (and expensive) insulation or by the use of "guard-ring" techniques. In view of all the factors concerned the best thing is to use a different

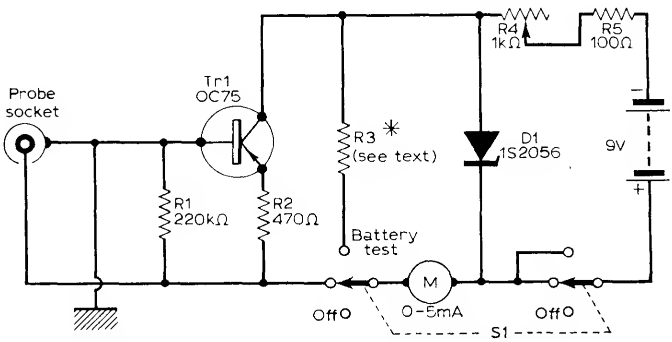


Fig. 1. The single-transistor circuit of the instrument.

method to achieve the required results. The device described here uses cheap and readily obtained components and good accuracy with complete safety results.

Fig. 1 shows the very simple circuit diagram of the instrument. It is best arranged as a small piece of copper-clad laminate and mounted in a tin box together with the meter and battery. The whole thing is then self-contained and may be connected to the chassis of a television receiver without any worries about battery polarity and so on.

The OC75 is chosen because of its high direct current gain, but because it is a *pnp* device and the e.h.t. of a television receiver is positive the circuit has to be arranged "upside down" as it were. Thus the meter battery and switch need to be well insulated from the containing box. The meter usually has a plastic case and the switch wafers (of ceramic or paxolin) are equally well insulated. The battery is a less reliable affair and it may be advisable to wrap it in a polythene bag before securing it in a suitable clip. A coaxial socket is provided for the input; the "inner" is therefore connected to the battery-positive line when the instrument is switched or while the base of the transistor is connected to the containing box. The latter must be of metal for safety and several types of household tins may be used.

If the d.c. gain of the transistor is 100, a change of base current of 50μA will give a collector current charge of 5mA. At zero input voltage, the base of the transistor is connected to the positive end of the emitter resistor (through a 220kΩ resistor) and the transistor is almost cut off. The slight leakage current is only a few microamps and reads on the meter. However, the deflection of the needle is so small as to be well within the zero adjustment.

When 50μA flows from the input terminal through the emitter-base-earth path the collector current is 5mA. Thus a robust meter can be used; a $2\frac{1}{2}$in diameter meter is quite suitable and may be obtained cheaply. If a 1mA meter is used, it may be shunted by a suitable resistor to give the correct full-scale deflection.

Construction of the Probe

The construction is straightforward but great care must be taken if safety is to be assured. The input current being 50μA maximum, for 20kV input a resistor chain amounting to 400MΩ in all is required, and this is made up to a close approximation by means of 18 22MΩ resistors in series with a 3·9MΩ resistor. Components of 20% tolerance can be used, since if these are all taken at random the errors tend to cancel out when the "sample" is as large as 18 or 19. If 10% tolerance components are used the total resistance will be within 2% of the nominal figure.

Using small carbon resistors of $\frac{3}{8}$in length and allowing a $\frac{1}{8}$in lead at each end for connecting up, the total length from the beginning of the first resistor body to the end of the 19th will be $8\frac{7}{8}$in. A piece of paxolin tube just over 9in will therefore be required and its diameter should be $\frac{3}{8}$in.

The 19 resistors are carefully soldered together in line, taking care to align them as accurately as possible; this is easy if the leads are bent over at a slight angle as close to the resistor body as possible. The chain can be rolled between two boards to finish off this operation (Fig. 2). Then to one end is soldered the inner insulated spill of a coaxial plug. The polythene type of plug is not so good as the p.t.f.e.; although the latter is rather more expensive, insulation is better and as the melting point of p.t.f.e. is much higher, soldering is simplified.

The resistor chain is next wiped over carefully with a *clean* rag soaked in carbon tetrachloride or lighter fuel to remove traces of grease and is then dried in a current of warm air. It is next passed through the paxolin tube and the polythene "inner" forced tightly into the end of the tube. The size is just right. The resistor chain is arranged to be quite central in the tube.

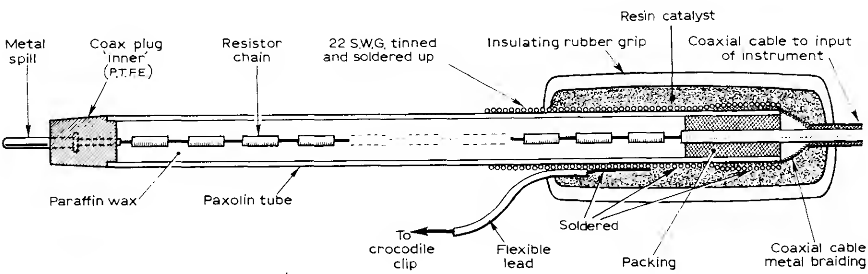
The next process is to melt sufficient paraffin wax (candle grease will *not* do) and to raise its temperature to drive off all air and moisture. This can be done in a small aluminium saucepan (preferably a good pourer!) and should be continued until the wax is well above the boiling point of water. It is allowed to cool a little and then the paxolin tube is filled up with the wax. On solidifying the wax contracts, so enough liquid should be available to "top up" the tube to above the body of the last resistor. When the wax is cold and hard the next step can be taken.

This is to solder the last resistor lead to the inner of a piece of coaxial cable and then to pack insulating material between the "inner" insulation and the paxolin tube. Strips of polythene are very good for this or alternatively insulation tape can be used. Lastly the outer braiding is teased out and slipped over the outside of the tube.

Then, starting at the end of the tube, 22 s.w.g. tinned copper wire is wound tightly over the braiding and one-third way up the tube. Finally this is soldered up to hold it in position and give good, solid contact all along its length. A bicycle type rubber or plastic handle grip may then be fitted over the end and filled with resin-catalyst (as used in car repair outfits) to strengthen

Fig. 2 (right). *The junctions between elements of the resistor-chain should be arranged as shown here.*

Fig. 3 (below). *A section through the probe showing its construction.*



the cable end. The complete assembly is shown in Fig. 3, which also shows how an earth lead is attached to the probe.

The purpose of the bare-wire winding is twofold. In the first place it protects the user, since if a flashover should occur (very unlikely) it will be intercepted before reaching the fingers. In the second place, any leakage that occurs will be along the outside of the tube, since the insulation inside the tube is entirely adequate. This leakage current will be directed to earth and not pass through the transistor and so will not affect the reading.

The cable may be of any convenient length and is terminated in a coaxial plug which fits into the corresponding socket. Whether the instrument is switched on or off no more than 11V can appear across this socket.

Calibration merely consists of passing a known current into the socket and noting the meter deflection. This is readily accomplished by means of a battery, a variable resistance and a 0–50μA meter. Since the total resistance of the probe is known quite accurately and the input resistance of the instrument is negligible compared with the probe resistance a simple calculation gives the value of e.h.t. which would produce the same input current. It has been assumed that transistor gain will be 100 but it may be considerably more or less than this and hence the full-scale deflection of the meter may correspond to other than 20kV. If the scale goes up to as much as 30kV it will be best to change the transistor for one of higher current gain. If the meter has been shunted to decrease its sensitivity, the shunt can, of course, be adjusted to obtain a full-scale deflection of around 20kV. Table A shows the values of e.h.t. corresponding to the calibration current used.

The scale will not be quite linear at the lowest readings but above 5kV linearity is quite good. There is an advantage in this in that the higher readings are spread over a larger proportion of the scale and are more accurately marked.

It will be noted that a zener diode is used together with a series resistor in the battery supply to regulate the voltage applied to the transistor. The variable resistor R4, also in series, enables an adjustment to be made for an ageing battery. The check is simple. The switch is turned to "battery test" position and R4 rotated from its minimum position until the voltage indicated no longer rises. At this point the zener diode is "in" and is regulating the supply. The value of R3 should be chosen in conjunction with the full-scale deflection of the meter used. The specified zener diode regulates at 5·6V, so if a meter deflection of 4mA approximately is required R3 should be 1·2kΩ.

The accuracy obtainable with this instrument is such that this simple regulation of the power supply is worthwhile for low e.h.t. readings especially. For readings above 10–15kV it is not really necessary but, if omitted, the battery-test position of the switch should be included to enable a check to be made from time to time.

In use, the resistor R4 should be adjusted so that the zener diode just conducts on test and no more. Higher settings merely waste battery current!

TABLE A

Calibration Current	*E.H.T. corresponding (mark the meter scale with this value)*
5μA	2kV
10μA	4kV
15μA	6kV
20μA	8kV
25μA	10kV
30μA	12kV
35μA	14kV
40μA	16kV
45μA	20kV

Chapter Twenty-four

A TEST UNIT FOR C.R. TUBES

THIS instrument tests older c.r. tubes while they are in position in the receiver, and enables faults to be found and the effects of possible remedies determined. When designing the unit, it was decided that it should have the following features:

1. Test for low emission, heater-to-cathode leakage, grid-to-cathode leakage, and low e.h.t.
2. Provision for boosting the heater voltage, firstly to check how much boost is needed to regain satisfactory emission, and secondly, to compare the picture with and without boost.
3. Provision for checking the effect of a low capacity isolation transformer on a tube with a heater-to-cathode leakage and for curing the usual type of grid-to-cathode leak.

Some of the above considerations tend to conflict with each other; for instance, if the e.h.t. were to be measured directly, either a bulky valve voltmeter would have to be incorporated, or a delicate microammeter movement used, and so this test is carried out indirectly.

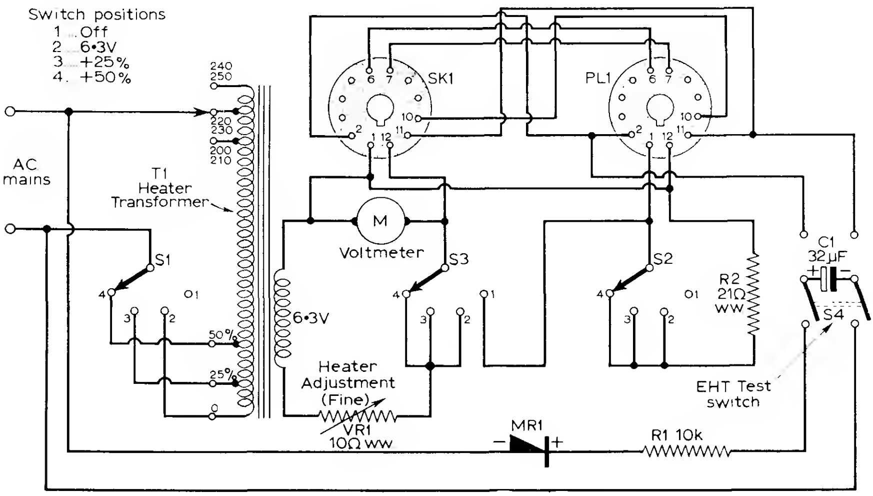


Fig. 1. The circuit of the instrument when using a mains transformer with a tapped primary winding.

Circuit

Since many tubes used the 12-pin base with standardised connections and 6·3V heaters, the unit is built using a 12-pin plug (an old c.r.t. base is suitable) and a 12-pin socket. If required, adaptors could be made for other tube bases and another transformer used with more voltage tappings.

The circuit of the test instrument is given in Fig. 1 and the plug and socket are wired up as shown with leads about 3ft long so that the instrument may be used while watching the picture on the c.r.t. The heater voltage for the c.r.t. is selected by means of S1/S2/S3 and VR1, and its value is shown by the a.c. voltmeter M. T1 is a low-capacity 6·3V transformer with provision for 25% and 50% boost to the heater voltage. In position 1 of S1/S2/S3, the heater of the tube is connected to its usual source of power (the heater chain in an a.c./d.c. set). In position 2 of S1/S2/S3, the heater is isolated from the set, but is not boosted, thus testing for heater-to-cathode leakage. In positions 3 and 4, 25% and 50% boost respectively are applied to the heater.

Switch S4 is a biased double-pole, change-over switch which is normally in the position shown in Fig. 1. C1 is charged to about 300V by the mains via MR1, but, on operating S4, the condenser is disconnected from the mains and connected between the grid and cathode of the tube. This puts a high positive bias on the c.r.t. grid, which, if the e.h.t. is at a workable level, should cause a bright flash on the screen. On releasing S4, the tube should re-light after a few seconds, and C1 is left to be charged up through R1 ready for use again.

The circuit in Fig. 1 uses a transformer in which the primary is tapped

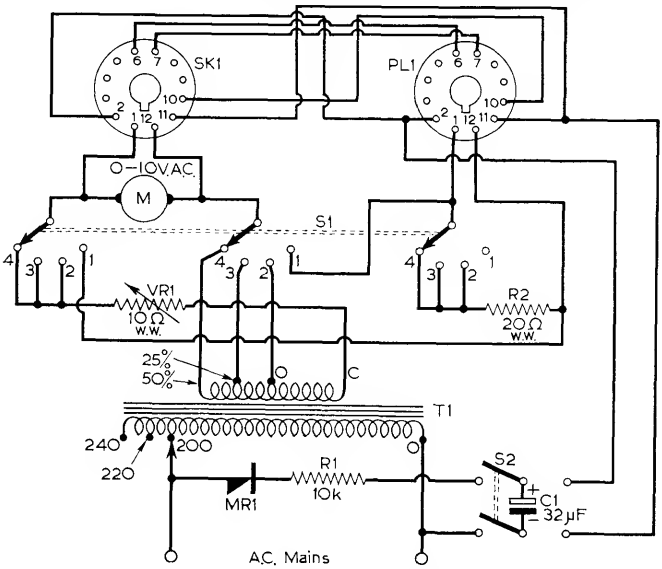


Fig. 2. An alternative circuit in which a transformer with a tapped secondary winding is used.

to provide boosted heater voltages; an alternative circuit using a transformer with a tapped secondary is given in Fig. 2.

Construction

The unit is constructed in a wooden case and the dimensions are given in Fig. 3. The wiring of the test instrument is shown in Fig. 4, and reference to this diagram and the illustrations will make the wiring easy to complete. The layout of the front panel is given in Fig. 5.

Testing a Tube

The plug PL1 is connected to the tube-socket in the set and the socket SK1 to the tube-base in the set and the test instrument plugged into the mains, and the TV set switched on. With S1 in position 1, the set is allowed to warm up. The meter M now reads the heater voltage of the tube; if it is zero, either no power is reaching the set or the heater chain is open circuited. If the needle is hard over, the heater of the tube is open circuited, although this fault would normally be found during preliminary checks of the receiver. Slight errors in the heater voltage are of little consequence but could be caused by the use of the incorrect mains tapping on the set.

The switch S1/S2/S3 is now turned to position 2 and the heater voltage set to 6·3 with VR1. Should the picture now be normal, the tube has a heater-to-cathode leak, and a low capacity isolating transformer will be required to

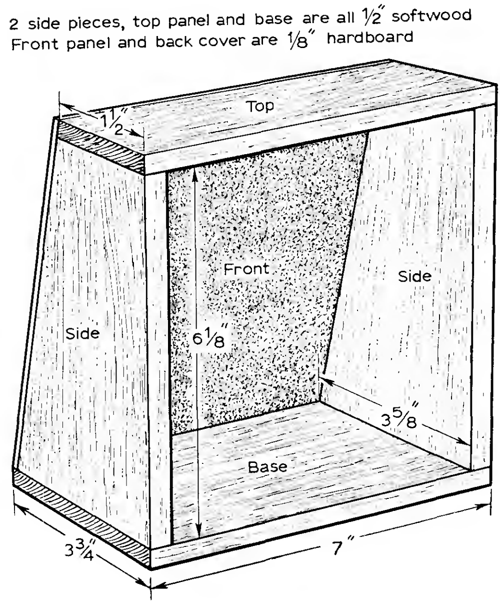


Fig. 3. The construction of the cabinet.

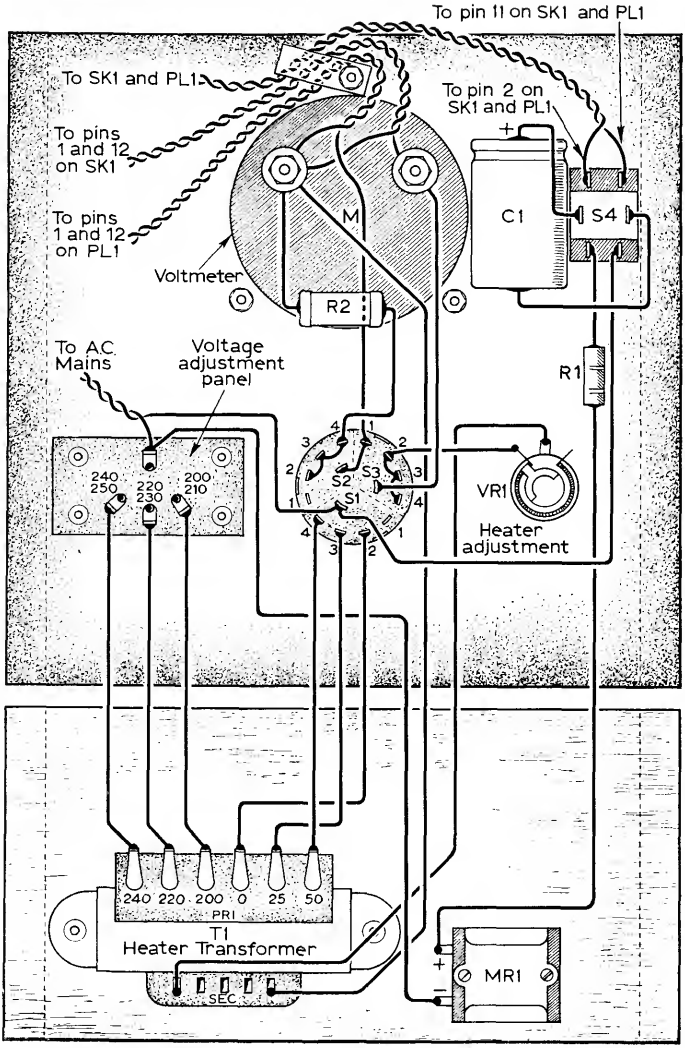


Fig. 4. The wiring of the unit.

cure the fault. However, if the picture remains poor, S1/S2/S3 must be set to position 3 or even position 4, adjusting VR1 to give an acceptable picture. The voltmeter M then shows the voltage needed from a boost transformer.

If the picture is still not good, the cause is usually either low e.h.t., or a grid-to-cathode leak. If, on pressing S4, a bright flash is seen on the screen of the tube, the e.h.t. is probably in order, and the fault may be a grid-to-cathode leak. These leaks are often caused by small pieces of conducting

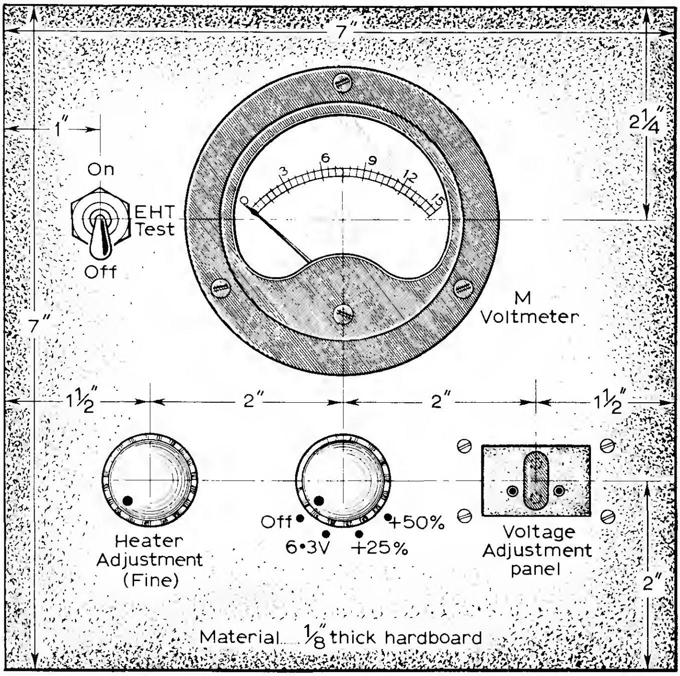


Fig. 5. The layout of the front panel.

material lodging between the electrodes of the tube, and the sudden application of a high voltage pulse often burns them away. This pulse is provided by the charge in C1.

The whole process of using this c.r.t. test instrument takes less than three minutes, and shows at once the effects of boosting and other processes without long delays and soldering of connections.

Cabinet

In order to make the instrument portable, it is as well to house it in a strong cabinet. The cabinet shown in the illustrations and in Fig. 3 was made from ½in softwood and ⅛in hardboard, the hardboard being used for the front panel and the back of the cabinet.

Construction of the cabinet should be begun by cutting out a piece of hardboard 7in square and marking it up for the actual components to be used in the instrument. The parts to be mounted on the front panel are indicated in Fig. 5. The meter will probably need a circular cut-out and a fret-saw or coping-saw should be employed for this operation.

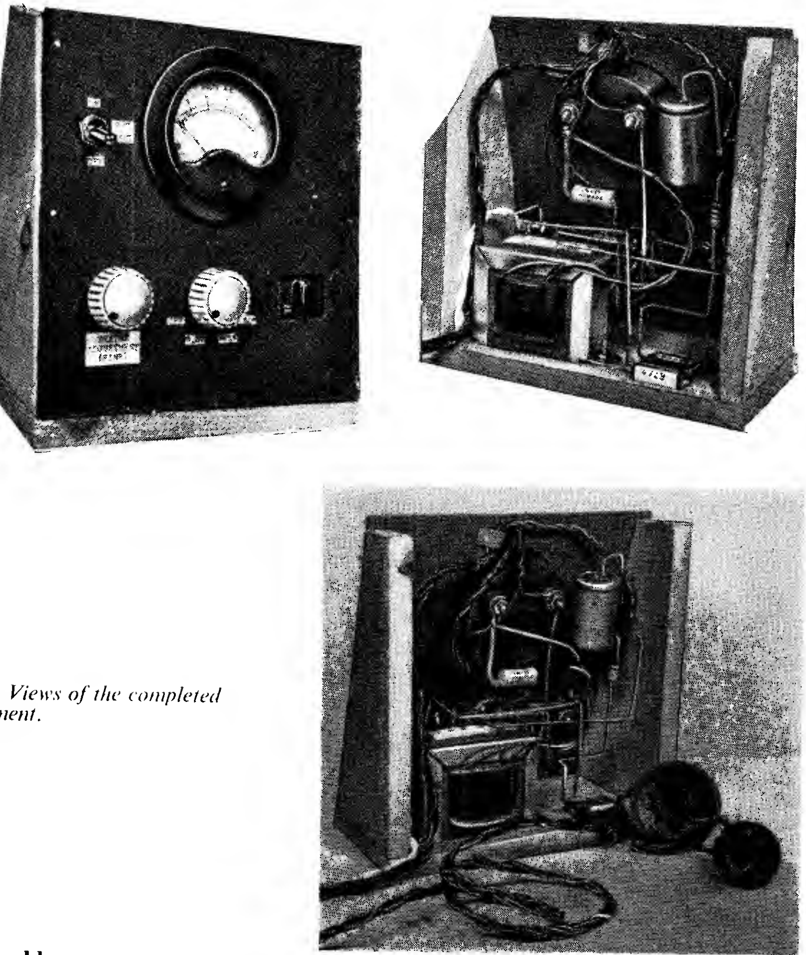


Fig. 6. Views of the completed instrument.

Assembly

The sides of the cabinet should be cut to the dimensions shown in Fig. 3 and if possible should be cut together so that they are both the same shape. The front edges of the base and top of the cabinet will need to be shaped to conform to the slope of the front panel.

Before assembling the cabinet, fit the necessary parts on to the front panel to ensure that they fit accurately. Then, remove them and use panel pins or screws to assemble the cabinet but do not fit the back, of course. When all the hammering has been completed, then the meter and other parts may be fixed in position of the front panel.

The back of the panel should be screwed on so that it may readily be removed if any of the parts of the unit need to be replaced.

List of Components

Resistors:
R1 10kΩ $\frac{1}{2}$W
R2 21Ω 3W wire-wound
VR1 10Ω wire-wound potentiometer

Capacitor:
C1 32μF 350V electrolytic

Switches:
S1/S2/S3 3-pole, 4-way
S4 Biased double-pole, change-over

Rectifier:
E250C50 or other small mains type

Transformer:
Low capacity c.r.t. boost type (see text)

Meter:
A.C. voltmeter: 0–10V or 0–20V or similar

Miscellaneous:
12-pin c.r.t. base and 12-pin socket; wire, solder, screws, wood for cabinet, etc.

Chapter Twenty-five

A SIGNAL-STRENGTH METER FOR AERIAL ADJUSTMENTS (BANDS I AND III)

MANY do-it-yourself television enthusiasts like to make their own aerials, but are presented with a problem when it comes to comparing one make or design with another. There is also the difficulty of orientating the finished array for the best possible signal pick-up. One way out of the latter difficulty might be to turn the set-connected aerial until the best picture is received; but this is a possible solution only with "vintage" receivers.

The majority of sets made during the last six to seven years feature reasonably effective automatic gain control (a.g.c.) on vision as well as on sound. This means that the picture and sound remain reasonably constant even though the signal applied to the set may rise and fall widely as the aerial is adjusted for direction.

As the aerial is turned and the signal rises, so the "noise" or grain on the picture reduces due to the improvement in signal-to-noise ratio. The picture does not usually become very much brighter; nor does it dim much as the aerial passes away from the direction of maximum signal pick-up.

Difficulty may thus be encountered by endeavouring to employ the set as a signal-level indicating device. If the set is stood by a window or even outside so that the man at the aerial can see the screen, a small change in picture noise just cannot be discerned, and a considerable error in aerial orientation is likely to result. Turning up the sound will not help either, for this will barely change in volume as the aerial is turned through a full 360°.

The professional aerial rigger overcomes these problems by the use of a signal-strength meter, which is an instrument capable of reading the actual signal strength at the end of the coaxial downlead on any particular channel —sound or vision. The instrument employs an ordinary moving-coil movement calibrated direct in millivolts and microvolts. Thus the aerial is connected to the input socket of the instrument which is then tuned to the correct channel and adjusted for maximum signal. The signal voltage is then revealed and it becomes a simple matter to turn the aerial to secure the maximum reading.

These instruments are rather expensive, and unless a lot of aerial work is proposed it would barely pay the average enthusiast to purchase one. However, it is to be shown here how an indicating meter may be fitted to any television set to record relative signal strengths, and a circuit will also be given for a signal-strength meter.

Basic Signal-Strength Meter

Before we go to these items we should understand briefly how the commercial type of signal-strength meter operates, for the same general principles will also apply to our equipment—as, indeed, to any equipment of this kind.

The block diagram in Fig. 1 gives the general idea. Here we have a tuner which selects the signal to be measured and converts it to an i.f. (intermediate frequency). This signal is fed to an i.f. amplifier, the anode current of which is monitored on a milliammeter. A "set-zero" control is also incorporated in the i.f. amplifier circuit so that with the aerial (or signal) removed from the tuner, the i.f. amplifier current can be adjusted to give exactly full-scale deflection on the movement. This deflection corresponds to zero signal.

Now the i.f. amplifier feeds a diode detector (rectifier) stage which is arranged so that the i.f. signal after rectification produces a negative potential across the detector load resistor. This negative potential is fed back as bias to the i.f. amplifier. Thus as the signal strength increases so the negative bias increases and the anode current of the i.f. amplifier valve *decreases*. This decrease in current from full-scale deflection as registered on the milliammeter indirectly corresponds to signal voltage, and the scale of the meter may be calibrated accordingly.

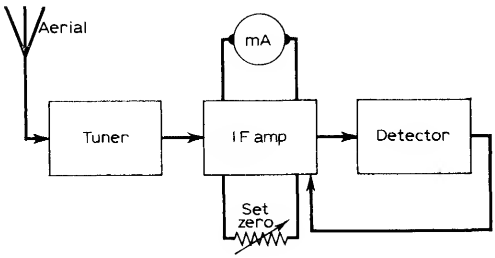


Fig. 1. The block diagram of a commercial signal-strength meter. The operation of the instrument is explained in the text.

Using a TV Receiver

It is quite easy to adapt any television set so that it also will give a similar indication of signal strength. In Fig. 2 is given the circuit of the common i.f. amplifier stage as found in many receivers. While there may be small differences in circuit and tuned-coupling detail, the essential features are always the same. The common stage receives i.f. signals from the tuner, and this stage invariably has a.g.c. applied to it. The valve may be an EF80 or similar pentode, instead of the EF85 of Fig. 2.

Movement Shunt

The anode current of the common i.f. amplifier valve with no a.g.c. bias applied is about 10mA, so the anode current meter should read, at least, up to this value of current f.s.d. The best idea is to employ a 0–1mA movement,

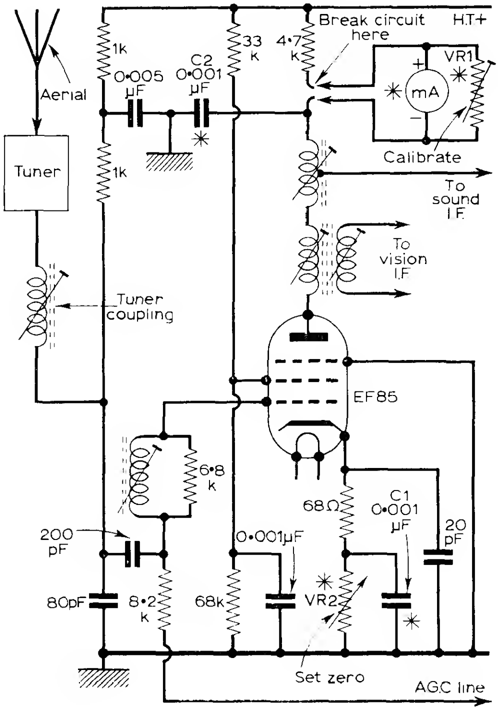


Fig. 2. The circuit of a common-i.f. stage in a typical television receiver showing how a milliammeter may be connected in the anode circuit to provide indications of relative signal strengths. Components marked with an asterisk are the extra ones required. Care should be taken with insulation to prevent shocks from a.c./d.c. receivers.

as these are readily available in all sizes and the movement can be shunted with a variable resistor as a means of calibration (VR1 in the circuit).

A 100Ω 0–1mA movement requires a shunt of 1·01Ω to give 0–10mA, and a 1Ω resistor shunted by a 10Ω pre-set have been found suitable by the author. Although there is very little full-scale variation given by the pre-set, sufficient is available for the purpose of the exercise. If possible, several 1Ω resistors should be measured for resistance and one with a slightly higher value than 1Ω should be selected.

Both of these components are available from Radiospares Ltd.—the 1Ω (rated at 3W) and the slider pre-set at 10Ω. (Note that Radiospares supply only to dealers, so the parts will have to be obtained from a dealer.)

If the movement has a resistance other than 100Ω the correct shunt value can be found from the following expression: Shunt in $\Omega = r/(n-1)$ where r is the internal resistance of the meter and n the ratio between the original and the new full-scale deflections. The whole idea, of course, is to secure exactly full-scale deflection on the meter when the aerial is removed from the set and with the contrast control full on.

Set Zero

Now when this condition has been established with the meter connected in the low signal-potential side of the anode load, a variable resistor should be connected in series with the cathode resistor of the stage (VR2 in the circuit). This will provide a main control of anode current and will allow the movement to be set to zero under a wider range of conditions (note that "set zero" means full-scale deflection—see the description of the signal strength meter in the earlier paragraphs). This resistor should have its spindle brought out to the rear or the side (or the front if required) of the set so that it is easily adjustable and its value should be 100Ω.

Connection from the cathode circuit to the variable resistor should be kept as short as possible and capacitor C1 should be included to rid the circuit of stray i.f. signals. The capacitor should be connected as close as possible to the valveholder and not to the control-end of the lead.

Meter Connection

The meter movement can either be mounted on the receiver cabinet somewhere convenient or it can be mounted in a small box so that it can stand on the top or close to the set and be connected to the set through a screened lead. If mounted externally the leads must be kept short and capacitor C2 employed. It is a good idea to include C2 even though the meter may be mounted on the set. *Great care must be taken with insulation* since most TV sets are of the a.c./d.c. type.

Not all sets will have the type of anode circuit shown in Fig. 2, but the idea is to connect the meter in the h.t. feed to the anode of the valve at the h.t. side of any coils that may form the anode load. The circuit, of course, is first broken, as shown in Fig. 2, and the gap is bridged by the meter movement with the positive side of the movement to the h.t. line side of the circuit. Decouple the movement at the *negative* side with C2.

Operation

After the set has properly warmed up the aerial should be removed and the contrast turned to maximum. The "set-zero" control should then be adjusted to give exactly full-scale deflection on the meter. When the aerial is connected, the reading will fall by an amount governed by the signal strength. A very strong signal will cause the meter to read almost zero current (owing to the resulting heavy negative a.g.c. bias), so it is impossible to damage the movement by too strong a signal.

Exact signal-strength calibration is possible if a signal generator is available with an accurate attenuator and which tunes over the television bands. The scale on the meter may be changed for one made out in microvolts or a graph can be drawn giving signal strength against meter reading.

Since old receivers can now be obtained for a few pounds the experimenter

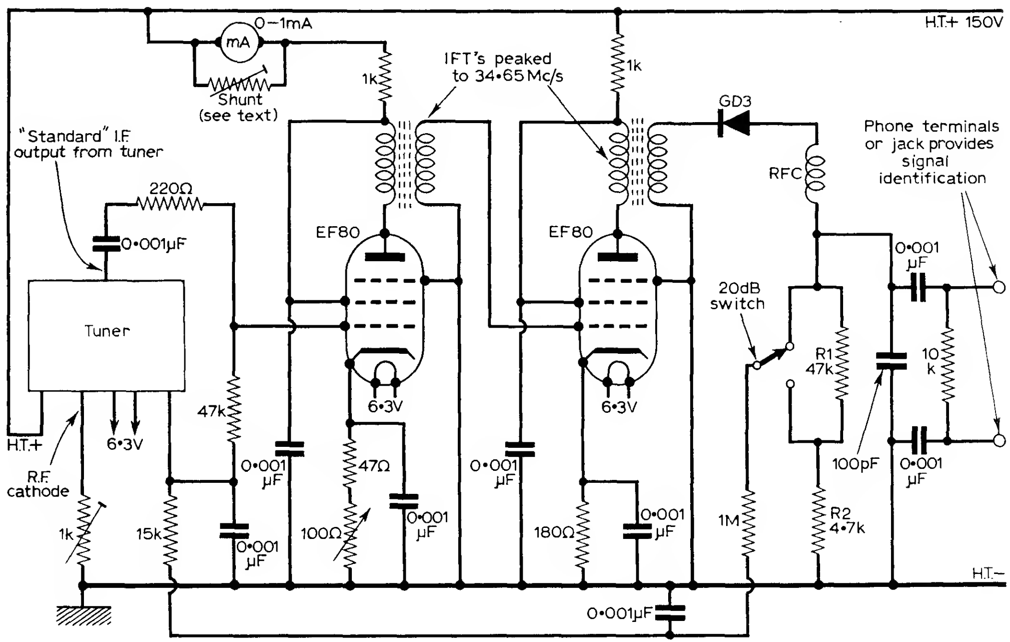


Fig. 3. The circuit of a single-strength meter for comparative tests. The milliameter may be calibrated direct in mV and μV or a graph may be produced giving signal voltages against meter readings. For u.h.f. signals in Bands IV and V, the tuner would cover the u.h.f. channels, and the a.g.c. would be removed from that section.

may consider it worthwhile to purchase a specimen solely to convert into a signal-strength meter. The sound and vision need not work properly provided the a.g.c. and common stage, including the tuner, are fully active. However, it is desirable to be able to monitor the sound and vision signals from the loudspeaker and on the tube as a check for interference and ghosting, which are factors that the meter movement is not able to disclose. Note that although the current in the common stage is metered it is only the vision signal that is measured, since the meter deflection is a function of the a.g.c. bias which is derived from the vision signal.

The set must, of course, have some sort of a.g.c. system, but if this is not present on vision, the current in the anode circuit of the sound i.f. stage may be measured (provided sound a.g.c. is employed). Here, though, there is less sensitivity and it is best to use the vision signal if at all possible. On a very old model one may consider it worthwhile to add vision a.g.c. from the sync separator control grid circuit if it is not featured on the set as it stands.

It will be found that the meter reading alters as the contrast control is adjusted. This is because on mean-level type of a.g.c. systems the negative bias produced by the signal at the sync separator is countered by a positive potential applied to the a.g.c. line from the h.t. line via the contrast control. This is of no consequence provided the contrast is first set before adjusting the "set zero" control.

Signal Strength Meter Circuit for Bands I and III

A circuit specifically for a signal strength meter is given in Fig. 3. Here a television tuner feeds two stages of i.f. amplification. These stages need not be broadly tuned as in the vision channel of a television set and for that reason considerable gain is possible with two stages. The first stage is the controlled one and the control bias is derived from a GD3 or equivalent germanium diode; it is also a good idea to take the control to the tuner as well to avoid overloading on very strong signals.

The switch S1 acts as a 10-times (20dB) attenuator; in the position shown the voltage across both R1 and R2 in series is used as control, while in the other position the voltage across R2 only is utilised—this, of course, is the 20dB position.

It will be seen that the circuit operates in exactly the same way as considered in the foregoing paragraphs. While a signal strength meter or comparator is highly desirable for aerial tests in Bands I and III (and the f.m. Band II) such an instrument is almost essential for Bands IV and V. Tuners for u.h.f. are readily available, so it is only a matter of making up a circuit similar to Fig. 3, with the i.f. transformers tuned to 39·5Mc/s.

Chapter Twenty-six

A TRANSISTORISED TEST OSCILLATOR FOR U.H.F.

THE design described here is an attempt to meet present-day needs in a modern manner.

This instrument is fully transistorised and is thus independent of domestic mains. There is, in fact, not a great deal of advantage in dry battery operation except that the "capital" expense of mains transformer, rectifier and smoothing components is avoided, because unless the matter is given proper attention loss of calibration as battery voltage drops in use can be a nuisance. However, the mains lead is a most effective distributor of stray radiation, especially at high frequencies, and to be able to dispense with this simplifies the quite thorny problem of preventing such leakage. For the home constructor battery operation at u.h.f. is particularly useful, since without precision machining facilities he can usually approach the high degree of screening needed.

Oscillator

The u.h.f. oscillator consists of a transistor arranged in a tuned-line Colpitts circuit, the common-base configuration being employed. This type of circuit is widely used in u.h.f. work whether transistors or vacuum tubes are utilised, since the inductive and capacitive circuit elements are much more manageable in regard to physical size than if any attempt were made to use very small coils and capacitors. Also the mechanical stability of this arrangement is inherently better, while losses tend to be lower if proper care is taken. In fact while a complete u.h.f. television tuner needs a specially made three or four gang capacitor—following an appropriate tuning "law" —this instrument needs only readily obtained components.

The transistor specified is the GMO290 by Texas Instruments, Manton Lane, Bedford. It is available at an early delivery date at a very reasonable price. This device is capable of oscillation at 1000Mc/s or even at a higher frequency, so no difficulty will be experienced in getting the circuit going.

Modulator

The modulator is arranged to supply line sync pulses only, although in fact irregularities in the switching waveform do actually provide something

more than a blank raster when the signal is picked up by a television receiver. The provision of a complete interlaced pattern, with both line and field signal pulses, would require much more comprehensive circuitry and it would cost much more to build the generator. While this might be desirable for the service engineer, the home experimenter will perhaps find the modulation afforded sufficient for his needs.

The modulator can be switched so as to give an audio signal instead of sync pulses and the frequency of this signal is a reasonable compromise between audibility and complexity of switching. While a little on the high-pitched side—about 2·5kc/s—it is at least distinctive!

Care has been taken in design that the audio modulation causes the least possible frequency modulation of the carrier. This may seem a little surprising at first; having in mind that the receiver circuitry is intended to receive frequency modulated transmissions it may seem odd to ensure that little or nothing is actually heard when the signal is received! However, it will be remembered that when an f.m. signal is detected the discriminator has to be adjusted so that amplitude modulation disappears. Tuning the detector consists in rotating the discriminator inductor core to get a sharp null point when an amplitude modulated signal is being received. Thus a signal generator is most useful for alignment if it is capable of providing an a.m. carrier with little or no frequency modulation superimposed.

Attenuator

The attenuator is of the cut-off waveguide type. The greatest difficulty exists in constructing an attenuator of the resistive type which will be useful at the ultra-high frequencies. Stray capacitances are a source of great trouble, since they offer reactances comparable with the value of the resistance itself. Also the physical size of resistors is not a negligible fraction of a wavelength; hence the voltage-current relationship along a resistor is not the same everywhere and its real value is very different from its d.c. or low-frequency value. In addition, and for the same reason, radiation is emitted and can be picked up elsewhere in the attenuator. While the exercise can be carried out it is not one for the man without a well-equipped laboratory!

The cut-off-waveguide attenuator neatly sidesteps these difficulties by not using resistors at all and by employing radiation as the means of transferring energy. In Fig. 1 the elements of this device, usually called a "piston attenuator", are shown. It consists simply of a hollow tube of a good conductor (usually copper or brass) whose diameter is small compared with a half-wavelength of energy to be transmitted. At 1000Mc/s the half-wavelength is 15cm and a tube some 1in in diameter or less would be suitable. At one end an insulated metal disc is connected to the source of energy, while separated from it is another insulated disc which receives energy, acting as a receiving aerial.

A TM_{01} wave is propagated down the tube from the launching disc and because the tube is too narrow to sustain such a wave the attenuation is

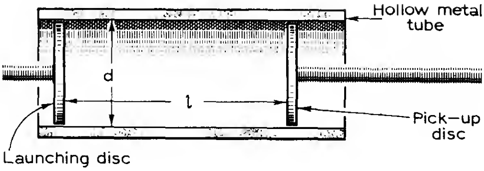


Fig. 1. Elements of the piston attenuator.

severe. The amount of energy picked up by the receiving disc is thus very small. However, the amount of attenuation in decibels is exactly dependent on the spacing of the two discs and if the receiving disc is movable its position will be on a linear decibel scale of attenuation. The relationship is so precise that anybody who has access to a good machine shop can make for himself a primary-standard attenuator working on this principle. The method used in this instrument is much less precise than this but nevertheless gives excellent results. Fig. 2 shows the circuit diagram and also indicates the way in which the attenuator is supplied with "input". The signal at the launching disc is measured by means of a u.h.f. diode and a microammeter with a series resistance—enabling it to read 200mV (full-scale deflection).

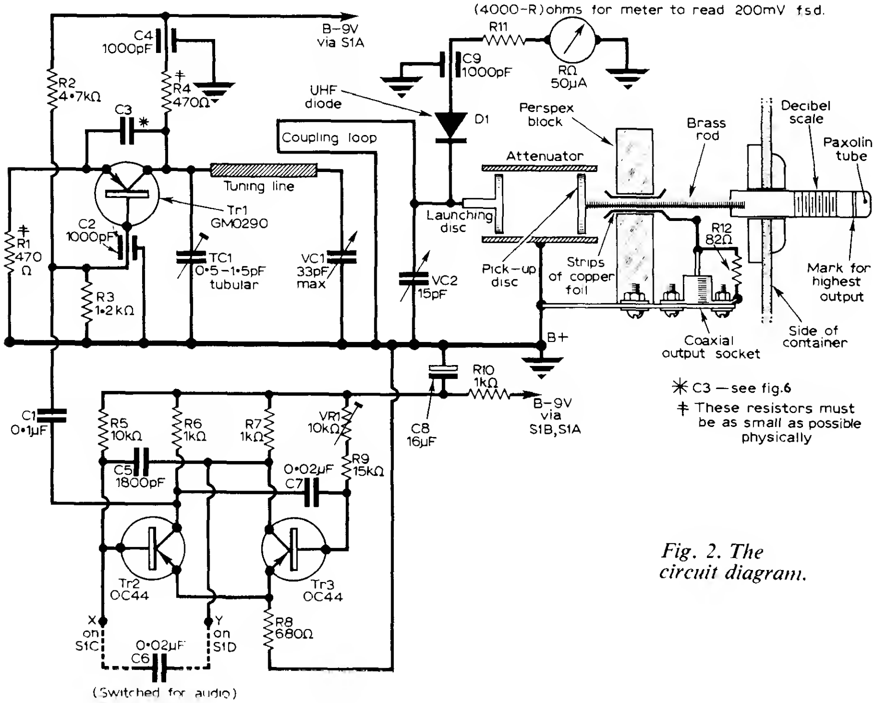


Fig. 2. The circuit diagram.

This meter is in a way optional, since the instrument will work without it. It does enable the attenuator input to be brought to a definite value, however, and thus for the instrument's output to be more meaningful. The adjustment of attenuator input is carried out by tuning the coupling loop, while the movable disc can be set to give a definite value of attenuation and hence a known signal in the output socket. It will be noted, however, that with the simple slide contact shown, even though the pick-up marker is terminated in an 82Ω resistor, there is always an unmatched rod of metal in the vicinity actually connected to the pick-up disc. This will give rise to errors in current distribution, since it has to be several centimetres long to work at all, and thus the attenuation scale will not be correct for all frequencies. The errors are not great provided the zero of the decibel scale is begun 5mm or so from the point where the two discs are actually touching. (This position gives the highest output at the coaxial socket.)

Wiring Techniques

It will be realised that good v.h.f. techniques need to be developed—and, as it were, pushed to the limit—when constructing a circuit for u.h.f.; even a few millimetres of component lead offer appreciable reactances at such frequencies. If when working on medium frequencies (say 1Mc/s) a lead length of 6in is tolerable, the comparable lead length at 1000Mc/s is under 1/100in! Conversely if for practical reasons leads must be (let us say) ¼in in length at u.h.f., the circuit may be expected to compare with a medium wave circuit using component leads over 20ft long. The constructor who has not yet made himself a thermal shunt (from a crocodile clip and a couple of inches of 12 s.w.g. copper wire) should do so as a preliminary so that, for example, transistor leads may be reduced to the minimum and still solder up without damage to the active transistor areas.

Figs. 3 and 4 show how the transistor is mounted so as to achieve minimum lead lengths. The whole u.h.f. circuit is contained in a trough of sheet aluminium measuring 10cm in length, 4cm in width and 3cm in depth (22 s.w.g.). A lid is folded for this trough out of another piece of sheet aluminium and is later secured in place by means of six self-tapping screws.

The tuning capacitor is a Jackson Brothers C804 50pF. The full 50pF is not required to achieve the tuning range and so to minimise self-capacitance and reduce the capacitance some of the rotor and stator plates are removed with a hacksaw, leaving five rotor and four stator plates. The sawing operation must be done carefully and lightly to avoid distorting the assembled component. In order to make the correct connection to the tuning line a small piece of copper foil is soldered to the lugs of the stator plates as shown in Fig. 5. Owing to the relatively large thermal capacity of the mounting pillars of the capacitor an instrument-type soldering iron alone will provide insufficient heat; either a heavier iron must be used or auxiliary heating given to the small iron, for example by heating in a gas flame. The tuning line consists of two tapering straight pieces of 22 s.w.g. tinned copper wire,

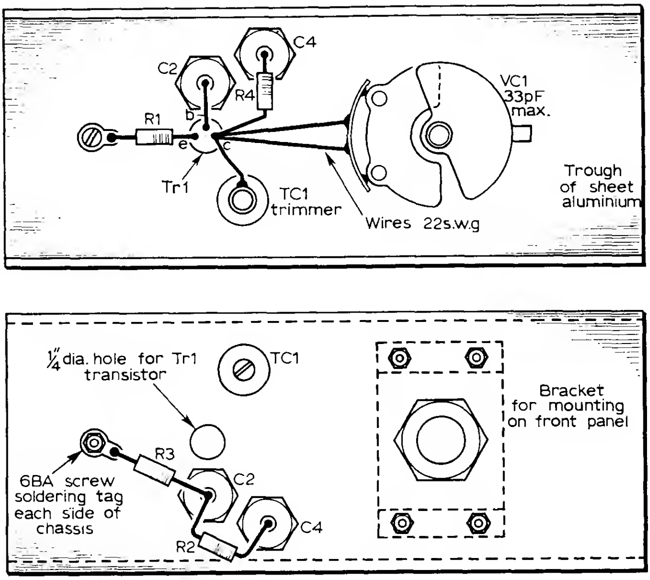


Fig. 3 (right). *The arrangements of the components for the oscillator.*

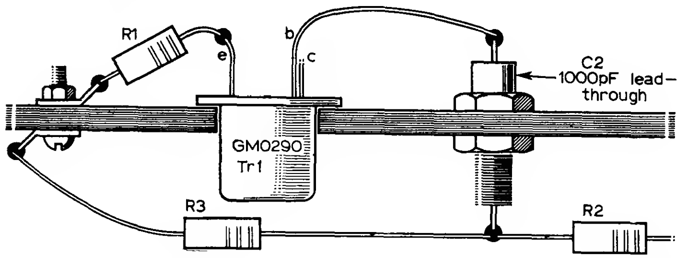


Fig. 4 (above). *Details of the mounting of the transistor Tr1* (*GMO290*) *with the shortest possible lengths of leads.*

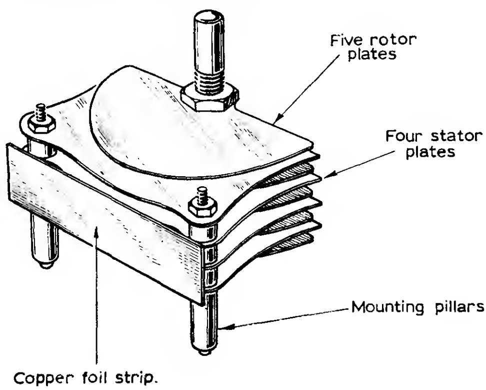


Fig. 5 (right). *Connections to the stator of the tuning capacitor.*

separated by $\frac{3}{16}$in at the tuning capacitor and joined together at the transistor end. The holes for the transistor and the tuning capacitor are 3·7cm apart and the wires should be cut to fit precisely into place between them.

The coupling capacitor which enables feedback to take place need only be very small. A component of the correct value (0·25pF) is not commonly stocked by shops but can be fabricated simply (as shown in Fig. 6) from short lengths of enamelled copper wire linked lightly together. A single link would be enough—it is not necessary to twist the wires together.

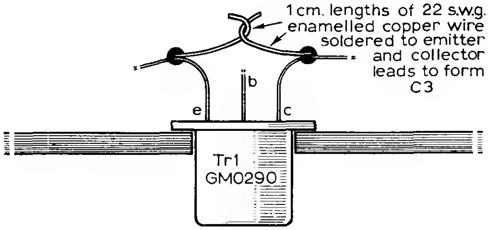


Fig. 6. Emitter-to-collector coupling; the fabrication of the coupling capacitor C3 (see Fig. 2).

Coupling Loop

The coupling loop which transfers energy from the oscillator circuit to the output socket is made up as follows: A length of p.v.c. covered connecting wire (not flex) measuring 15cm in length is folded in half and twisted together lightly for a short portion of its length—about 3cm. The remaining portion is fashioned into a rectangular loop of sides 3cm and end 1·5cm. The loop is positioned above the tuning wires so as to afford reasonable pick-up of energy; this is best adjusted later when the u.h.f. diode and meter are connected. Without untwisting the remainder of the wire the ends are separated and bared and are soldered to stand-off insulators in appropriate positions. Connections are then made by strips of copper foil or heavy flex to the attenuator and adjusting capacitor VC2 (Jackson type C804, 15pF).

When the above assembly has been completed and before the lid is placed in position the oscillator circuit should be adjusted to the correct tuning range as follows: Set VC1 to its minimum position and connect the power supplies. Check (for example with a TV receiver) that Channel 68 can now be tuned by very slight adjustment of VC1. If not and the assembly is correct a fault in wiring should be looked for whereby the minimum capacitance in circuit is too large. If none appears on inspection the length of the tuning line will have to be reduced and VC1 repositioned. Next mesh the plates of VC1 fully and adjust until Channel 21 is reached. Again adjust VC1 to its minimum and check that Channel 68 can still be tuned. Now adjust the position of the coupling loop until a reasonable reading (say 10μA) can be obtained on any channel. Note that VC2 will need alteration to adjust the meter reading—its effect on frequency is negligible. When the loop has been put into the right position a further check of maximum and minimum frequency should be made as loop position has a small effect on the tuning.

Diode

The u.h.f. diode used will depend on what is available; an ordinary OA70 will not do very well and several advertisers offer microwave diodes at low cost. These are often of the coaxial type and, if so, need small clips to hold them into the circuit. Such clips can be made of short lengths of thin brass sheet such as can be obtained from a spent cycle headlamp battery.

The attenuator tube in the prototype was a piece of brass tubing $\frac{9}{16}$in diameter and 2·2cm long. It was mounted by soldering on to it, centrally, a 4B.A. bolt. This is better than a clip as it allows for adjustment of length as needed. This gives an attenuation of 30dB/cm movement of the slide and, allowing for not starting the decibel scale at the point where the launching and pick-up discs are touching, a maximum attenuation of about 50dB. Two millimetres movement halves the output (-6dB) very closely, and marks 2mm apart on the instrument attenuator are not too close for good visibility.

If a tube of different diameter is used the attenuation can be worked out by remembering that it is universally proportional to the diameter. Thus a cigar tube of the usual aluminium type $\frac{3}{4}$in diameter will give an alternative attenuation of 22·5dB/cm.

Care must be taken to ensure that both discs are well insulated from the attenuator tube. The method of making electrical connection to the sliding disc will be obvious from Fig. 2. The Perspex block through which the guide hole for the shaft is drilled should be about 1cm thick and can be made up of several pieces of sheet Perspex cemented together with a drop of chloroform. It may be secured to the chassis by means of a 4B.A. bolt screwed into a hole drilled and tapped in the block.

If the capacitor VC2 is adjusted so that the meter reads 5μA, the minimum output from the attenuator is approximately 30μV with the two discs 2cm apart. If a smaller output still is required it will suffice to plug into the coaxial socket centre a short length of wire and use it as an aerial. This should be done in the final stages of lining up a u.h.f. receiver and it is also useful when checking the instrument's tuning range and modulation with a TV receiver.

Modulator Circuit

The modulator must comprise an asymmetrical astable multivibrator arranged to give the correct line scan and flyback period to a good approximation. If frequency modulation of the carrier is to be avoided a good square wave at 15,625c/s must be generated and this is accomplished by using high-frequency transistors type OC44 in a suitable circuit. The suggested printed circuit diagram is given in Fig. 7 or a small piece of Veroboard could be used. This multivibrator can be switched to a lower frequency, with almost symmetrical characteristics, for use as an audio modulator. Details of the function switch arrangements are given in Fig. 8.

The whole of the circuitry is attached to a piece of aluminium sheet which forms the front panel of a soldered-up tin of suitable size. This tin, crackle-enamelled for a good appearance, also houses the PP9 battery which powers

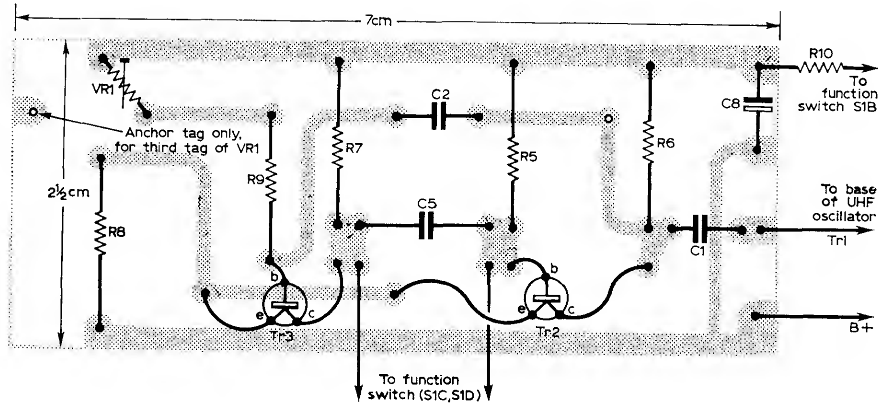


Fig. 7 (above). *A printed circuit for the modulator.*

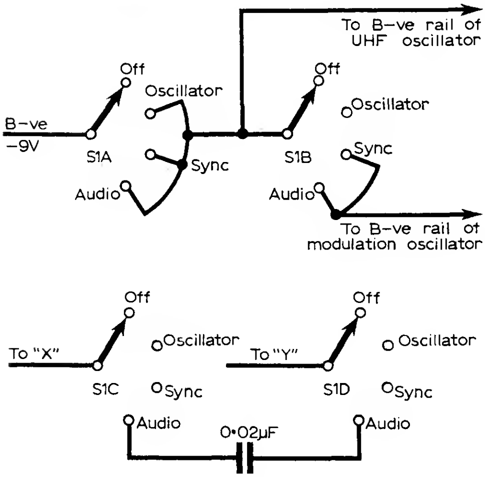


Fig. 8 (left). *Details of the function-switch.*

the unit. The front panel is secured in place by about a dozen self-tapping screws. The attenuator handle projects from the side of the box and the hole through which it passes is bushed to afford some support for the handle. A suitable bush can be made from an old potentiometer cut down to size and will naturally improve the appearance of the instrument.

Drive

A good slow-motion drive is essential for this instrument and suitable types are widely advertised. The prototype utilises a Muirhead epicyclic drive.

Calibration may readily be done by using a TV receiver with u.h.f. tuner. If this is not available, measurement of standing waves on parallel transmission lines (Lecher lines) gives very good results so long as the wires are accurately parallel and not too far apart (an inch or less). This needs patience, of course, but results to within 0·5Mc/s are not too difficult to achieve.

List of Components

R1	470Ω
R2	4·7kΩ
R3	1·2kΩ
R4	470Ω
R5	10kΩ
R6	1kΩ
R7	1kΩ
R8	680Ω
R9	15kΩ
R10	1kΩ
R11	See text
R12	82Ω

Potentiometers:

VR1 10kΩ pre-set

Capacitors:

C1	0·1μF
C2	1000pF feed-through
C3	See text and Fig. 2
C4	1000pF feed-through
C5	1800pF
C6	0·02μF
C7	0·02μF
C8	16μF or 25μF 25V working
C9	1000pF feed-through
TC1	0·5–1·5pF tubular
VC1	33pF Jackson Brothers (50pF) Type C804 modified. See text
VC2	15pF Jackson Brothers—type C804

Attenuator:

Brass rod. Perspex. Copper foil. Paxolin tube. Copper or brass tube 1in dia. Copper or brass discs

Miscellaneous:

Meter 0–50μA optional, see text

D1 Diode 0A70 (see text)

Transistor type GM0290 Texas instruments

Switch S1A, B, C, Yaxley 4P, 4W, 2B

Printed circuit board or piece of Vero board coaxial socket

Aluminium for front panel and trough

Chapter Twenty-seven

A WIDE-BAND OSCILLOSCOPE

In any serious development work with television circuits it is almost mandatory to possess an oscilloscope of reasonably high quality. Just how good it needs to be is not too difficult to establish and this will be dealt with later.

However, the demands of any television system are known to be somewhat severe and it must be expected that the simpler types of instrument will be inadequate except for the more rudimentary aspects of testing. The present design is intended to put within the reach of the amateur a high-grade oscilloscope capable of undertaking both more elementary and more advanced work and at the same time be inexpensive and relatively easy to construct both mechanically and electrically. A block diagram of the oscilloscope is shown in Fig. 1, and the specification of the instrument appears on page 368.

The Display Tube

To begin with, the design is centred on the old cathode-ray tube VCR97. This was at one time a favourite tube for experimental television receivers and proved to be a good picture tube of its kind. It is still available, as a look through current advertisements will show, and is cheap. However, there are few serious experimenters who do not have one laid aside somewhere.

There is, of course, no harm and indeed considerable benefit in using a more modern tube if desired. In any case a 5in or 6in tube is recommended. Although much good work can be done with tubes as small as even 1in diameter, the convenience of a 6in display is such that the average constructor will be convinced at first sight.

Oddly enough, the present British 405-line system presents the severest condition in displaying line synchronising pulses. The r.f. picture waveform is seldom if ever required for display except for its primary purpose of producing a picture. In this system the rise time is less than 0·25μs, while in the 625-line system a rise time of 0·256μs is specified. The difference is not great, however.

Rise Time

The rise time is usually given as the time taken for the edge of a pulse to rise (or fall) from 10% to 90% of its extreme values. The frequency response corresponding to this is given by the expression:

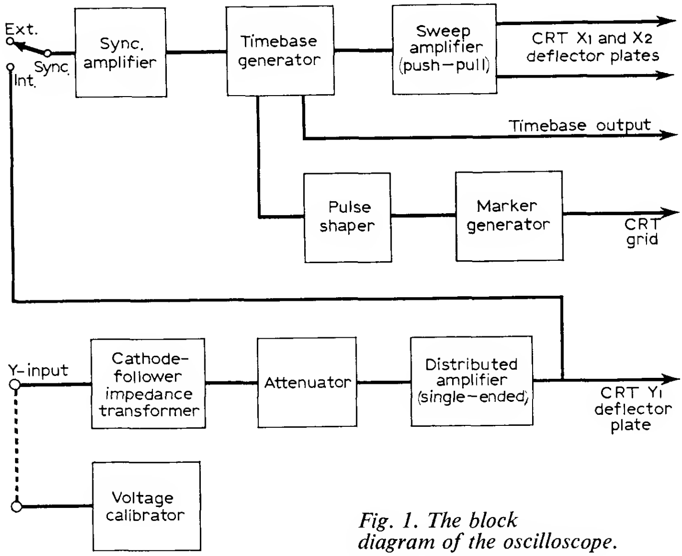


Fig. 1. The block diagram of the oscilloscope.

$$\text{rise time in microseconds} = \frac{0{\cdot}7 \text{ to } 0{\cdot}9}{\text{bandwidth in Mc/s}}$$

Thus a conservatively estimated bandwidth for the 405-line pulses would be 3Mc/s, which is, of course, the allotted bandwidth transmitted.

If, however, an oscilloscope is used whose Y-amplifier bandwidth is 3Mc/s, the Y-amplifier is also contributing an error; this amplifier would, of course, display an instantaneous pulse as one whose rise time was 0·25μs.

If a pulse of actual rise time 0·25μs were displayed it would appear as if it had a rise time of 0·35μs—an error of some magnitude. While this would hardly matter in routine receiver servicing, development work calls for something better in order of magnitude.

In the instrument here described the Y-amplifier is arranged to have a bandwidth of 20Mc/s. Although this bandwidth can be achieved by conventional amplifiers, here a "distributed" amplifier is used as it has several practical advantages, especially in its relative simplicity and stability; in addition the use of such an amplifier overcomes the difficulty of arranging an output stage capable of developing 80V or more to feed the Y-deflection plates of the cathode-ray tube. The design details will be discussed later.

The Timebase (see Fig. 4)

The timebase generator is of the familiar Miller-transitron type, modified to obtain extremely fast flyback.

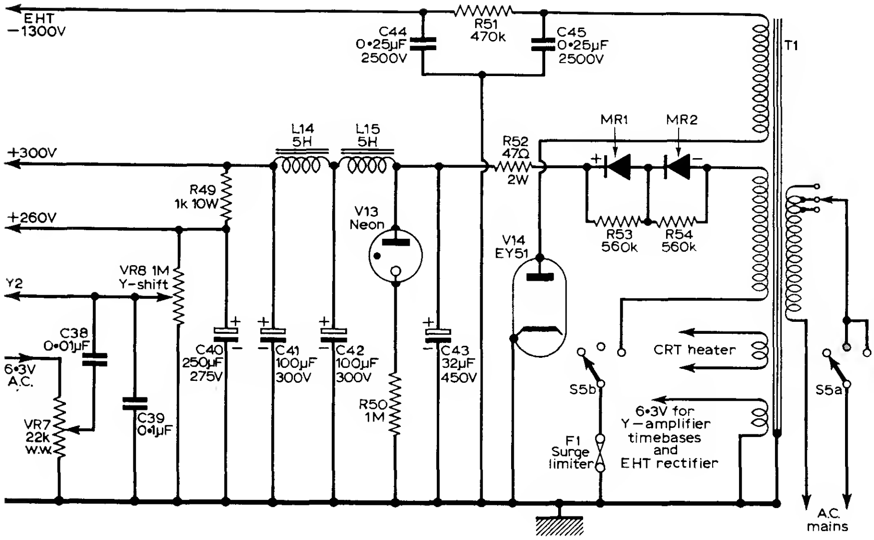


Fig. 2 (above). *The circuit of the power supply.*

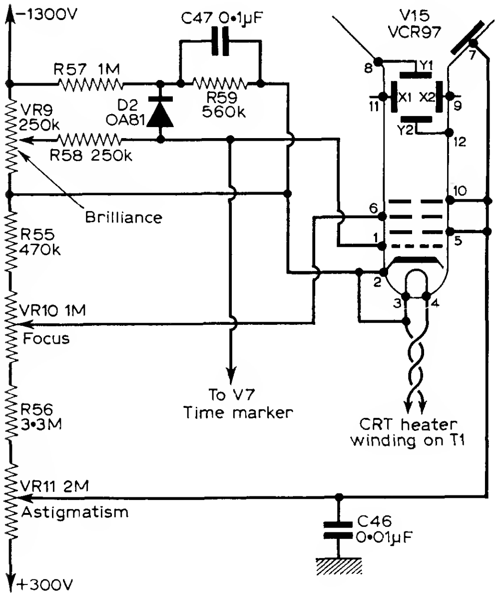


Fig. 3 (left). *The power-supply network for the c.r.t.*

Provision is not made in this circuit for flyback blanking of the cathode-ray tube display, although it can be incorporated, if desired, by a simple addition to the circuit. The reason for this apparent omission is that while at low speeds of scan the flyback pulse falls accurately on the flyback, at high speeds delay occurs in charging up the stray capacitances through the stray inductances of the circuit, and the flyback pulse is a fraction of a microsecond late in arriving at the cathode-ray tube.

Thus the first part of the trace is lost and maybe with it part of the pulse which ought to be displayed. Also, the time of flyback is not clearly defined, and so in measurements involving time or frequency, errors may occur because one or more cycles of the displayed waveform are lost in the flyback.

In this instrument the brilliance can be advanced to show the flyback trace and any part of the display so "lost" can be accounted for.

At the highest speeds the flyback trace takes an appreciable fraction of the scan time and is visible at all but the lowest brilliance setting. When being used on the first ten ranges, however, flyback is extremely rapid and although visible in a darkened room at fairly high brilliance, the effect is hardly noticeable at all in use.

Synchronisation

The timebase generator is synchronised either from an external source or by switching the Y-amplifier output to the sync amplifier. This consists of a pentode valve operated with a low and variable screen voltage. By this means the grid base is much shortened, so causing the valve to operate as a slicer. The actual grid base is determined by d.c. control of the screen voltage, which enables the signal to be kept out of long leads and also provides a measure of automatic synchronisation.

A positive-going signal on the grid of V1 causes the anode voltage to drop considerably; this negative pulse, if of sufficient magnitude to overcome the bias across V2, is passed to the anode of V3 and thence by way of the Miller capacitance to the grid of V3, cutting off the valve and allowing the anode voltage to rise to the level of the h.t. line. Thus the flyback is initiated.

The timebase generator produces a very linear sweep of about 40V–60V magnitude. This is not great enough to scan the cathode-ray tube and so an X-amplifier of the cathode-coupled triode configuration is employed to produce the necessary sweep voltage. This type of amplifier has the advantage that it is very insensitive to hum voltages and, although the power supply smoothing is of excellent characteristics, any help in this direction is of value.

Another advantage is that a push-pull output is obtained and this is of great value in avoiding deflection defocusing of the trace. With the VCR97 specified, good focusing can be achieved with asymmetrical deflection, but many modern tubes are sensitive in this respect.

Bandwidth

The bandwidth of the X-amplifier need not be very great and the 47kΩ

resistors in each anode need cause no surprise. With triodes, relatively high values of anode resistor, relative to the anode impedance of the valve, must be employed to achieve acceptable linearity of scan.

For this reason the ECC84 is used in this position: the anode impedance is low, even when the effects of the 2·2kΩ cathode resistors—which introduce negative current feedback—are taken into account.

The loss of bandwidth is not severe and in fact the trace is exceptionally linear except that on the highest ranges the corners of the sawtooth are rounded off, causing a little cramping in the first and last ¼in of the trace; the main part of the trace is still very linear.

X-Expansion

Since the timebase voltage is ample, the provision for X-expansion is readily provided. The trace can be expanded on all ranges to at least eight times the screen diameter, and expansion is from the centre of the trace, which proves to be most convenient in practice. The differential X-amplifier also allows of very ready X-shift by varying the grid voltage and hence the differential anode voltages.

Marker Generator

The screen pulse of the Miller valve, on flyback, is used to control the operation of the time-marker oscillator. The direct pulse is of the wrong polarity and is insufficient in amplitude, besides being of a rather indeterminate shape which additionally alters with the position of the range switch. Hence this pulse is shaped and amplified before being put into use.

This is accomplished by means of the pentode V6, which is operated with low anode and screen voltages and is direct coupled to the screen of V3. The 5pF capacitor across the 3·3MΩ coupling resistor is for "speed-up" purposes and its value was determined empirically to give the best pulse shape on the fastest timebase ranges, where the problem is most severe.

A good square negative pulse is hardly ever obtained on the grid of V6, but the positive pulse at its anode is well formed and of high amplitude.

V7A and B are connected in an oscillator circuit whose frequency can be switched to any of three spot frequencies, 1Mc/s, 100kc/s and 10kc/s. The oscillator is arranged to operate very "fiercely" and an output waveform is obtained which is rich in harmonics and is therefore a distorted sinewave, somewhat "spiky" in appearance.

This is just what is needed to modulate the cathode-ray tube beam so that the display shows, instead of a continuous trace, a series of sharply determined dots whose distance apart is 1μs, 10μs or 100μs as desired.

For switching on this oscillator at the beginning of each scan a special circuit is used which ensures that the oscillator always begins in the same phase. If it did not the dots would fall in random positions on subsequent traces and no markers would be visible.

Arrangements are also made for the grid of the cathode-ray tube to be kept below zero volts relative to its cathode: if this were not done the life of the tube would be shortened seriously.

This type of marker is generally preferable to that which causes a "pip" to be superimposed on the waveform displayed. This latter can often obscure small features of the wave, a state of affairs easier to avoid if the markers are of brightness-modulation. Brightness-modulation for this purpose is sometimes described as "two-axis" time marking.

Voltage Calibrator

As all three positions are needed for the time-markers, two of the remaining positions of the switch are devoted to the provision of 50-cycle calibrating voltages. These are obtained by means of resistive voltage dividers across one of the 6·3V transformer windings; 1V and 0·1V peak values are provided at a socket brought out to the front panel of the instrument, whence a small piece of wire only is needed to connect to the Y-amplifier input.

This is done to avoid internal switching, which necessarily has to be in circuit all the time and adds materially to the input capacitance of the Y-amplifier. In this circuit the Y-amplifier has a very high input resistance—over 1MΩ—shunted by only 8pF, and additional strays of even only a few pF are to be avoided if the care taken in design and construction is not to be nullified.

Y-Amplifier (see Fig. 5)

A stepped attenuator is provided, in preference to a continuously variable gain control, for the Y-amplifier, largely because the resetting accuracy is so much better. During development of the oscilloscope the attenuator was given more thought than might be imagined.

It was realised that if time-consuming and delicate adjustments were to be avoided the attenuator used would have to rely upon inherent independence of frequency rather than upon corrections to be applied after construction.

As is well known, at the higher frequencies the self-capacitances of resistors alter their effective values and a simple voltage-dividing network divides voltage as much by capacitance as by resistance. This can be avoided by using low-value resistors of a few tens or hundreds of ohms, but, of course, if such an attenuator were placed in the grid circuit of a valve the Y-amplifier would have an unacceptably low input impedance.

The case of the "few ohms" resistor, whose errors are due to self-inductance, can, of course, be neglected.

Accordingly, in this circuit, a cathode-follower input valve is used whose bias is given correctly by the cathode resistor employed. The output impedance is very low, suitable for feeding into an attenuator comprising

low-value resistors. The attenuator is terminated by the cathode impedance of a grounded-grid amplifier, the anode circuit of which develops the voltage which is fed into the grid line of the distributed amplifier proper.

This type of attenuator arrangement suffers from the disadvantage that the input cathode follower has less than unity amplification and that only half this reduced output can be developed across the following grounded-grid amplifier input. However, this reduction in gain can be made up later, and the circuit shows negligible frequency-dependence up to 20Mc/s, the bandwidth of the Y-amplifier.

Actually, separate tests show that negligible errors occur up to at least 40Mc/s, so there is plenty in hand.

Power Supplies (Fig. 2)

Power supplies for this oscilliscope are modest (and if the specified transformer is not obtainable a 350V–0–350V 100mA type could probably be pressed into service). The e.h.t. is derived from a half-wave 900V winding, giving 1270V rectified output at low load. The main h.t. supply is 300V at less than 100mA, and as the final anode of the cathode-ray tube is connected to the positive h.t. rail a total e.h.t. of 1570V is available.

Although the VCR97 is rated at 2000V, and can actually operate at 2500V to advantage, 1570V permits of a spot diameter of less than 0·5mm at good brilliance. If a 350V–0–350V transformer is used, e.h.t. will total only 1340V, but this will result in only minor degradation of the display. In this case some slight alterations in the resistor chain may be needed to obtain the correct range of focus adjustment.

P.D.A.

If a modern tube incorporating post-deflection acceleration is used a voltage-doubling arrangement raising the e.h.t. to a total of 2800V can be employed. The intermediate anode should be connected to chassis and the p.d.a. anode to the main h.t. rail; this gives a little post-deflection acceleration, but not very much, and permits normal operation of the tube.

A tube permitting asymmetric Y-deflection should be chosen, as the distributed-amplifier output is single-ended. Alternatively, a secondary-emission output valve can be added to the amplifier, push-pull output being taken from anode and dynode.

Anode and dynode resistance of about 2·2kΩ each would enable an appropriate output to be obtained, with a bandwidth of about 30Mc/s.

Construction

The instrument is constructed in double-decker fashion. The lower chassis is approximately 17in × 7in and the upper somewhat smaller, about 11in × 7in.

The lower deck is devoted to the larger hardware, such as the smoothing chokes and capacitors, and also the transformer, if a stout mumetal box is available for encasing it. If not, this item is best separated completely from the oscilloscope, appropriate leads being made up into cable form to connect the "power box" with the main instrument.

The importance of this can hardly be overstressed. The transformer field is certain to distort the trace appreciably unless very effective shielding is provided, and the usual mumetal screen around the cathode-ray tube is not normally effective enough to overcome the trouble.

It will be convenient to commence construction of this oscilloscope by installing the heavier and bulkier components on the lower deck and then completing the wiring of the power supply circuit. When this operation is finished, work on the upper chassis can be started. Consequently the portion of the instrument to be first described in detail will be the power section, and this will be followed by the timebase circuits.

Mains Transformers

The mains transformer T1 should have the following secondary windings: 0—300V 100mA, 0—900V 5mA, 6·3V 2A and 6·3V (tapped at 4V) 1A. Since this second l.t. winding supplies the c.r.t. heater, double insulation is desirable here. The primary should be shielded from the secondary winding by an electrostatic screen.

The 2A l.t. winding is used to supply the heaters of valves V1 to V6 and V14.

The Y-amplifier valves V8—V12 are supplied from a small auxiliary heater transformer T2. This is also used to provide the voltage-calibrating waveform, for reasons to be discussed later. It should be noted that this heater transformer is not shown in the circuit diagram (Fig. 2). However, no difficulty should arise on this account, as the actual wiring arrangement is clearly indicated in subsequent diagrams.

The heater of the cathode-ray tube VCR97 requires 4V at 1A, and if only a 6·3V winding is available a series resistor of 2·3Ω will be required. This may be made from a short length of Eureka resistance wire, wound on a ceramic former and fitted with fairly heavy-gauge copper wire leads. This assembly should be placed inside a Pyrex test-tube or similar housing to afford good insulation—it will be remembered that this series resistor is at a potential of 1300V relative to the chassis.

Arrangement of Components

Fig. 2 shows the circuit diagram of the power supplies (except for T2). If a mumetal box is available for containing the mains transformers T1 and T2 these, together with the rectifiers, chokes and resistive voltage droppers, are mounted on the bottom deck of the double chassis. This puts weight at

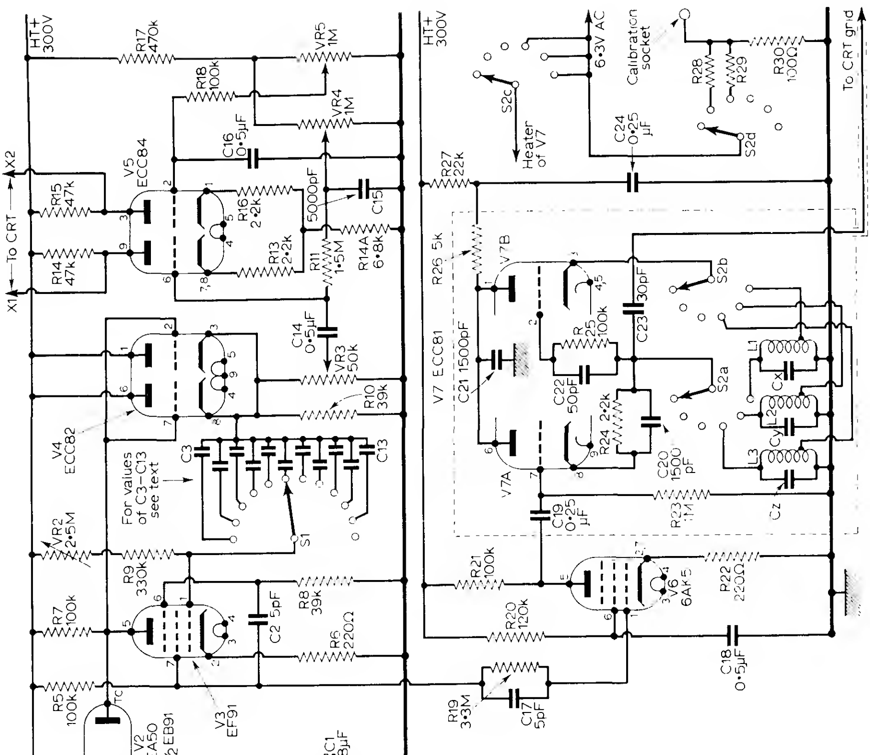


Fig. 4. The circuit of the timebase generator and amplifier, and the sync amplifier and time marker. The final "way" of S1 is connected to chassis—see text and Fig. 16. Note also that C13 is made up of "strays".

the base of the oscilloscope and makes it very stable in use. One practical point in construction is to place the heavier components symmetrically so that the carrying handle can be fixed centrally—otherwise a very odd appearance can result!

The power supplies are not stabilised, although this would be very desirable in general terms. However, since both X and Y axes are calibrated

(time and voltage respectively) there is little lost by the use of simpler circuits, and the added cost and extra heat produced are avoided.

The only thermionic valve used in the power supplies is an EY51 high-voltage rectifier (V14). Since a negative e.h.t. is required the heater of this valve may be run in parallel with those of the time-base generator valves, and so earthed to chassis.

Smoothing of the e.h.t. supply is performed by two high-voltage capacitors (C44, C45) and a 470kΩ resistor (R51) of low rating. Both C44 and C45 have a capacitance of 0·25μF and carry a high charge; a shock from these components can be extremely unpleasant and even dangerous, *so during construction the power should never be switched on for tests unless the cathode-ray tube chain of bleeder resistances is connected* (Fig. 6). This resistance chain will discharge the capacitors to a safe voltage a few seconds after switching off. Care is still needed, and the testing of this section of the apparatus should always be a leisurely affair, with plenty of time for thought before each operation. The e.h.t. of -1300V is connected to the junction of VR9 and R57 in Fig. 4.

H.T. Smoothing

The timebase generator and associated circuits need 300V at about 30mA, and the smoothing has to be really excellent. Elaborate arrangements are therefore made for reducing ripple on the h.t. supply: two chokes (L14, L15) are used with three capacitors (C40, C41, C42) in a double smoothing circuit which reduces ripple to less than 1mV.

The chokes should be preferably of the low-resistance type as used in television receivers, and they need to be rated for about 100mA since the Y-amplifier current is also supplied through them.

The rectification of the 300V supply is carried out by a pair of silicon diodes in series (MR1, MR2). Because with these rectifiers there is no warming-up period, it is necessary to delay switching on the h.t. until the valves are warmed up ready to take current; otherwise about 450V would be applied to the smoothing capacitors on switching on power. With the arrangement of Fig. 2 capacitors of 300V rating can be used without risk, and these are very much cheaper and more readily available than higher voltage types. The actual value of the smoothing capacitors is not critical, but larger values should not be used; the reservoir capacitor (C43) should be as specified.

The silicon rectifiers need protection in two ways from current surges. In the first place R52 limits the current flowing into the smoothing section—of which the most important here is the reservoir capacitor C43. On first switching on h.t. this capacitor acts as a dead short across the rectifiers and the current flowing is limited only by the total series resistance. R52 needs to be of adequate rating since heavy currents flow in it. The initial surge current is well over 1A, but this is within the surge current rating for 500mA silicon diodes.

In the second place there is always the possibility that an electrolytic capacitor may fail catastrophically, and if this happens the current flowing may rise sufficiently to destroy the rectifiers. For this reason a fuse is inserted in the common earth return of the h.t. transformer winding. A 6V or 12V 0·5A motor parking light bulb is used, since this will pass the normal current surges, which last only for a fraction of a second, while it will "blow" if an excessive current flows for one or two seconds.

Two Series-connected Diodes

The silicon diodes specified are rated at 1000V p.i.v., and it might be thought that a single one would do. This is definitely not recommended. The smoothing section L15, C42, L14, C41 is virtually a lumped-parameter delay line whose effective impedance is about 230Ω. On first switching on, C40 represents an initial "dead short" and the line is thus terminated in R49, 1000Ω. The sudden application of 300V h.t. therefore causes a pulse to be sent down the line towards the termination, and since the line is terminated in a resistance higher than its effective impedance a reflection of the pulse, without change of sign, takes place at the termination. A pulse of about 250V then comes back along the line, eventually arriving at the rectifier. The worst probable situation at the silicon diode is then as follows:

Voltage across C43		300V
Negative swing of transformer = $300\sqrt{2}$	=	425V
Pulse returning from smoothing circuit		250V
		975V

This is only a fraction below the maximum rating for the diode and leaves nothing in hand for transformer winding tolerances, mains surges, noise voltages on the mains and so on. Two such diodes are therefore specified, and to equalise peak inverse voltages each is shunted by a 560kΩ resistor of low wattage rating (R53, R54).

A small panel-mounting neon indicator (V13) is used as a pilot light to show when h.t. is switched on.

Hum-balancing Voltage

The small circuit at the extreme left of Fig. 2 is interesting in that an adjustable small alternating voltage is impressed on one of the Y-deflectors of the cathode-ray tube, as well as the variable d.c. potential which provides Y-shift. The reason for this is the fact that the valves of the Y-amplifier (to be described later) are r.f. valves and are therefore not especially well-suited to low-frequency amplification in that the hum induced from the heater is not minimised in manufacture to the extent to which such precautions are taken with (for example) the audio pentode EF86. Consequently, however well-smoothed the h.t. supply, there is always a few millivolts of hum in the output, and this is independent of the gain setting of the Y-amplifier.

By taking a tapping on a 6·3V winding by means of the potentiometer VR7, and applying this hum voltage to the other Y-plate of the cathode-ray tube, the effects of the two alternating voltages can be made to cancel very accurately. The effect is best seen, and the adjustment made, with a very low timebase speed synchronised with the mains, It is very much better if the waveform of the hum and of the correcting voltage are sinusoidal, because then there are no worries about phase differences between components of the two waveforms. It is best therefore to feed the Y-amplifier, and derive the correcting voltage, from the auxiliary heater transformer. The presence of rectifier pulses in the main transformer causes quite severe distortion in the waveforms found even in the heater windings. For this reason the voltage calibrating waveform also is taken from this separate heater transformer.

The 260V high tension supply is for the Y-amplifier. The EF184 valves need an anode voltage of about 180V, but the extra 80V is dropped across the anode load resistors.

Voltage Divider Chain (Fig. 3)

When wiring up the voltage divider chain for the cathode-ray tube, the network D2, R57, R59 and C47 should be added. The purpose of this, which will be explained fully later, is to protect the tube when beam-modulation is used; its purpose will not be self-evident during preliminary tests.

The chain of resistors shown in Fig. 3 is correct for the VCR97 but may not be very suitable if other types of tube are used. In particular, the "position" in the chain of VR10 may need to be altered; this may be accomplished easily enough by a little experiment with the value of R56 (up or down in value), as the focus requirements of the substitute tube demand. It may be noted that the resistor chain takes about 250μA while the cathode-ray tube anode takes a negligible current. Consequently tube current need not be considered in estimating the "position" of VR10. However, brilliance and focus are interdependent to a small extent, and both are dependent on the position of the slider of VR11, the astigmatism control.

In this circuit VR11 controls the voltage of the final anode of the c.r.t. and is used to optimise the latter in relation to the mean potential of the deflector plates. In practice there is no difficulty in securing a well-focused trace at any reasonable brightness by the simultaneous adjustment of focus and astigmatism controls. Thereafter there is no need to alter the astigmatism controls except when brightness is altered materially, one of the shift controls is used to effect a very great change in trace position, or a very large scan is used for some special purpose.

Timebase Generator

The timebase generator section of this instrument comprises the valves V1 to V5 (see Fig. 4). V1 is the sync amplifier, and for this application

the valve 6AK5 (EF95) has been chosen. It has not only the advantage of low inter-electrode capacitances, but is also economical of heater current. VR1 enables the screen voltage, and thus the amplification of the valve, to be varied. The screen is decoupled by the electrolytic capacitor C1, and thus a simple d.c. control of synchronisation voltage is obtained.

It is especially important to avoid r.f. in the control leads wherever possible as all the controls have to be brought out by way of fairly lengthy leads to the front panel. Owing to the relatively high anode load resistance of 23·5kΩ—the resistors R1 and R2 are in parallel with respect to the anode signal—the amplification rises with decrease of signal frequency, and this helps to compensate for the very small capacitance used to couple V1 grid to the Y-amplifier. Useful amplification is obtained up to a frequency of at least 25Mc/s and sync at the highest frequencies is easy to obtain.

Internal or External Sync

Either internal sync from the Y-amplifier, or externally applied sync, is available. Selection of the service required is made by means of a single-pole changeover switch S3. Here the presence on the front panel of a component carrying r.f. can hardly be avoided, and to keep the leads as short as possible it is arranged that V1 is as near as may be to the front panel. Priority in position has however to be given to another circuit (to be dealt with later) in which short leads are even more vital.

V3 is the Miller-transitron sweep generator, and V4 is a cathode follower so arranged as to give a very rapid flyback. Coupling between screen and suppressor of V3 is by way of a fixed 5pF capacitor (C2), and this proves sufficient to give reliable oscillation over all the frequency ranges covered. While this provision simplifies switching it does result in a very limited voltage sweep. The sweep varies a little from one range to another, but is about 40V—enough for the purpose required since push-pull amplification is arranged in the following stages V5a and V5b.

The valve EF91 is not intended for use as a suppressor-gated valve, and it might be thought preferable to use one designed for the job, such as the 6F33. The 6F33 works well in this circuit, but needs a larger screen-suppressor coupling capacitor of about 22pF or a little less. Twelve samples of the EF91 have been tried out in the prototype and have given consistent results, so the constructor can use this type with every confidence.

Rapid Flyback

The cathode follower V4 consists of both halves in parallel of the valve ECC82 and it speeds up flyback in the following way. In the simpler Miller transitron-sweep generator the timing capacitor—one of the series C3 to C13—is connected between anode and grid of the valve, and a high resistance connects the grid of the valve to h.t.

During flyback this capacitor has to discharge, and it does so via the anode resistor and the grid resistor in series. Because the latter is normally quite high the time of discharge is relatively slow. In this circuit, when the timing capacitor has to discharge it does so through the output impedance of the valve V4, which is 1/gm in parallel with the cathode resistor and amounts to about 200Ω, and the grid-cathode resistance of V3 under grid-current conditions—about 1000Ω. The total resistance in series with the capacitor is about 1400Ω, and discharge is extremely rapid. Flyback thus occupies only a small fraction of the sweep, even at the highest frequencies it occupies a quite negligible proportion of the sweep.

A very fast flyback means that except at the highest frequencies the return trace is invisible unless brightness is much increased. Consequently it has not been found necessary to include flyback blanking of the trace. On the two highest frequency ranges the return trace is visible, but at much reduced intensity. It is considered an advantage that in fact no part of the complete cycle is irretrievably lost, especially when the number of cycles displayed is to be counted. If desired, however, a small capacitor may be used to couple the anode of V6 to the cathode-ray tube grid, when complete blanking may be achieved. The capacitor used must have an ample voltage rating, and a suitable component can be made by twisting together lightly a few inches of p.v.c. covered connecting wire. However, owing to the presence of distributed inductance and capacitance in the circuit the flyback pulse is delayed in arrival at the cathode-ray tube grid, and the first part of the trace itself is lost. This may not be considered a disadvantage, as there is ample scan in any case.

Scan Speed

The prototype instrument has eleven ranges of frequency; within each range the scan speed is controlled by VR2, a 2·5MΩ linear-law potentiometer. The values of timing capacitor are as follows:

		Sweeps per second (approximately)
C3	1μF	2 to 10
C4	0·5μF	5 to 25
C5	0·15μF	12 to 50
C6	0·05μF	50 to 250
C7	0·01μF	200 to 1000
C8	2000pF	1000 to 5000
C9	500pF	4000 to 20,000
C10	200pF	6000 to 30,000
C11	50pF	20,000 to 60,000
C12	22pF	40,000 to 200,000
C13	strays	150,000 to 500.000

The extra contact on the switch S1 is connected to chassis direct. When this position is selected the timebase generator is out of action and an external timebase generator may be used if desired. A socket is provided for the timebase generator output, connected to the slider of VR3, and this does duty also as the input for an external timebase voltage. For this application the slider should be turned as for maximum output and the external generator —if of high impedance—should be connected to the socket via a cathode follower. The input impedance at this terminal is about 200Ω.

Differential Amplifier

The timebase amplifier V5 is a differential amplifier of the "long-tailed pair" type. As well as providing push-pull deflection voltages for a single input, an X-shift potential exists at the anodes which can be varied by altering the d.c. potential of one of the grids by means of the potentiometer VR5.

VR4 varies the timebase voltage fed to V5a grid and this is valuable for expanding the X-deflection. The un-bypassed cathode resistors R13 and R16 virtually lengthen the grid base of the valves to about 5V. As the timebase output is about 40V an eight-fold expansion is available on most ranges. The required part of the waveform displayed can be brought to the centre of the tube face by means of the Y-shift control VR5, providing the degree of expansion is not too great.

Physical Layout

The circuits comprising the sync amplifier, timebase generator and sweep amplifier are built along one side of the upper chassis so as to bring the X-amplifier anodes as close as possible to the base of the cathode-ray tube. This avoids excessive circuit capacitance and minimises the length of leads which at the higher timebase speeds carry high-frequency currents and may give rise to unwanted coupling between X and Y plates of the tube.

The leads and controls so far discussed and brought out to the front panel are as follows:

VR1	...... sync amplification	*VR11	...... astigmatism
VR2	...... X speed	S1	...... X range
VR3	...... X expansion	S3	...... sync switch
*VR5	...... X shift	*S5	...... power supplies
*VR9	...... brilliance	Coaxial socket	...... X output
*VR10	...... focus		

Those marked with an asterisk (*) are d.c. or mains-frequency items and the length of lead is unimportant. The remainder carry timebase frequency currents and the length of lead should be as small as possible. They should be

unshielded to minimise circuit capacitance but should consist of twisted flex to carry "go" and "return" currents. The exception to this is the case of the leads to the sync switch, where even the self-capacitance of twisted flex is unacceptable. The front panel should not be relied upon for an earth return except for the d.c. controls. The remaining panel controls are concerned with time and voltage calibration and the Y-amplifier. These will now be discussed.

Marker Oscillator

Along the same side of the upper chassis is built the marker oscillator (V7). Because this generates high frequencies at considerable voltage its necessary controls on the front panel must at all costs be connected by the shortest possible leads. Thus the valve nearest the front panel on this side is associated with this oscillator circuit.

In Fig. 4 the portion of the circuit around V7 is shown enclosed in a screening can and in the prototype it was found possible to construct the whole marker oscillator, including valve socket, on the lid of a small tin (actually a salted peanut tin) secured to the chassis. The tin itself then formed the removable screen. The most thorough isolation of this circuit from the rest of the wiring proved necessary and hence not only is the circuit screened but all the leads for power supplies—including heater wiring—as well.

The oscillatory voltage from this unit, used to modulate the cathode-ray tube beam current, is carried to the tube grid through a section of coaxial cable; the outer braid is earthed at only one point, as close to the oscillator circuit as possible and actually on the valve socket itself. Certain leads cannot, of course, be screened effectively; one of these is the wire carrying a timebase pulse from the anode of V6, where the capacitance of even a short length of coaxial cable would distort the pulse unacceptably. Here a short lead is necessary, well separated from any other connecting wires.

The operation of the circuit is as follows: V7b is a cathode-Hartley oscillator whose frequency of operation is determined by the value of inductance (L1 to L3) switched into circuit by means of S2a and S2b. The valve V7a is a cathode follower whose cathode resistor (2·2kΩ shunted by a 1500pF capacitor) is attached direct to the "live" end of the tuning inductor in circuit. If the grid of V7a is at such a potential that the valve draws anode current, cathode-follower action takes place across the tuning inductance and very heavy damping is imposed on it—sufficient to prevent oscillation. Also the "live" end of the tuning inductor is held at very near earth potential.

On applying a positive pulse to the grid of V6 a negative pulse of increased amplitude appears at the grid of V7a. This cuts off the valve V7a and the damping resistance in its cathode circuit is immediately removed. Oscillation now begins with the transfer of energy from the inductor to its tuning capacitor on the first half-cycle. Oscillations thus begin with practically full amplitude and always start in the same phase.

If this were not arranged and the oscillator allowed to start in random

phase, the markers produced would not occupy the same position on each trace; except by luck they would not be seen at all. The same would apply in the case of a continuously-running oscillator whose output to the cathode-ray tube was "gated" by a timebase pulse.

Pulse Shaper

The pulse shaper and amplifier V6 is direct-coupled to the screen of the Miller sweep generator whose screen generates a negative pulse on flyback. The capacitor C17 ensures only that the steep leading edge of the pulse is not lost, while the d.c. connection provides that when there is no screen pulse the grid of V6 is clamped to very nearly earth potential. The anode and screen load resistors are such that during the screen pulse V6 is cut off and this ensures that a pulse of constant amplitude is passed to the grid of V7a.

Because V7b is driven hard its output waveform is not sinusoidal but very peaky—just the sort of wave needed for use as a time marker. On arrival at the grid of the cathode-ray tube the varying potential causes a variation in beam current which in turn alters the brightness of the trace. Since the marker oscillator always starts at the beginning of the trace in the same phase the trace appears to be composed of a series of dark and light patches; the bright dots represent the tips of the modulating wave and are therefore separated by the time interval corresponding to one cycle of the wave. In Fig. 3 the diode D2 is so arranged as to conduct when the grid potential approaches that of the cathode. With the components specified the diode conducts before the grid voltage becomes positive and so the tube is well protected from damage.

Inductor Details

The inductors L1 and L2 are wound on the same former to conserve space. This is a bakelite tube 2·4in in length and $\frac{9}{64}$in diameter—the inner former of a popular type of canned assembly widely obtainable.

L1, the 1Mc/s oscillator coil, consists of 150 turns of 40 s.w.g. enamelled copper wire in three layers of 53, 50 and 47 turns respectively. The cathode tap is at 53 turns from the earthy end of the winding. A dust-iron slug is used for trimming and a fixed capacitor of 50pF for tuning is connected across the whole coil. This winding is placed near the bottom of the former.

L2, the 100kc/s oscillator coil, consists of 1600 turns of the same wire, pile-wound between cheeks $\frac{5}{8}$in apart. This winding is tapped at 300 turns from the earth end. A tuning slug and a capacitor of 200pF are provided. L2 should be situated on the former as far away from L1 as possible.

For L3, the inductor for 10kc/s, a different technique has to be used in order to economise in space. A few laminations from an old loudspeaker transformer were selected; they measured $2\frac{1}{4}$in $\times$ $\frac{3}{8}$in $\times$ 0·32in. Five such strips

were cut into three pieces of equal size except for one of double length which was drilled with a $\frac{1}{8}$in hole and bent over at right-angles for mounting purposes. Thus a stack of 14 laminations about $\frac{3}{4}$in $\times$ $\frac{1}{2}$in $\times$ $\frac{3}{8}$in was obtained.

On this stack was mounted a pair of end cheeks, the whole was dipped in a solution of shellac in methylated spirit and allowed to dry thoroughly (this took over 24 hours) before winding the coil.

Then 1600 turns of 40 s.w.g. enamelled copper wire were wound on, tapped at 400 turns from the earthy end for the cathode tap. With the laminations used, a capacitor of 0·001μF was found to give a close approximation to the frequency required, with a final adjustment (arrived at by trial and error) by an extra capacitor, nominally 82pF, in parallel.

The constructor will need to experiment a little in this circuit. The keen ear can arrive at precisely the television line frequency of 10,125c/s by direct comparison with a synchronised receiver. This is only $1\frac{1}{4}$% in error and may be sufficiently accurate.

If the ear is not sensitive enough a synchronised TV receiver can be used in any case, by feeding a little of the oscillator voltage direct to the cathode of the picture tube and adjusting for zero beat as seen on the tube face. This calibration is the most difficult in practice and the full drill for all ranges will be given in detail later.

Properly speaking, the Y-amplifier comprises the valves V9, V10, V11 and V12 with their associated circuits. However, V8 also carries the Y-signal, so it will be convenient to treat this whole series as one (Fig. 5).

In order to provide gain control the problem of resetting accuracy has to be solved. In this instrument continuously variable gain control has been avoided, partly because it is very difficult to make such a control independent of frequency, but chiefly to allow the use of a stepped attenuator whose resetting error can be nil.

The usual type of input attenuator is a frequency-compensated network consisting of a ladder of resistances and capacitances. Resistors can be selected from stock to high accuracy, or 1% tolerance components can be bought. The compensating capacitors however have to be adjusted, after the attenuator has been constructed, until the attenuator is independent of frequency. This can be a very patience-consuming operation and may take several days of spare-time work. It was decided to use a simpler method.

Resistors of relatively high value—20kΩ and upwards—are increasingly frequency-dependent in value (the Boella effect), due most probably to inherent self-capacitance effects. Resistors of some 100Ω and below are increasingly frequency-dependent because of their self-inductance. In the range of about 100Ω to 2000Ω the reactive effects are generally negligible up to around 50Mc/s at least, providing that good quality carbon composition resistors are used, without a spiral track. The problem of frequency-independent attenuators then resolves itself into a method of using this range of resistors to obtain the necessary attenuation, having regard to the fact that the impedance of a Y-amplifier as a whole must be high and as non-reactive as possible.

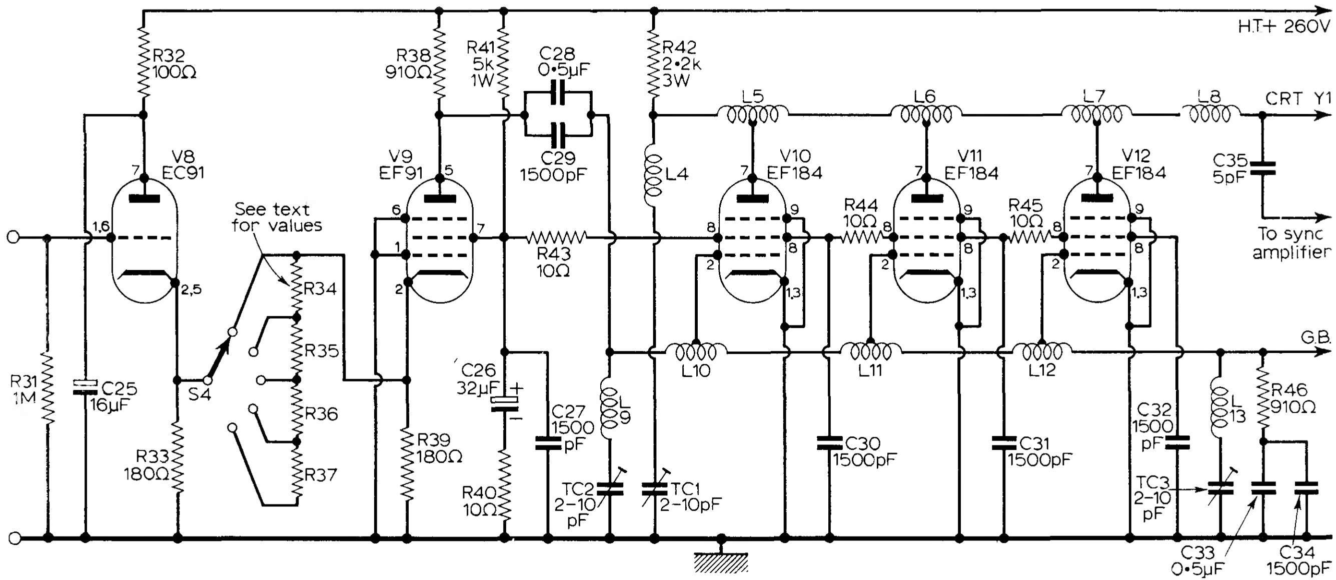


Fig. 5. The circuit of the attenuator and Y-amplifier. R33 and R39 should be 175Ω and 180Ω respectively, within 1%. They may be selected from resistors of nominal value 180Ω.

Cathode-follower Input

This is accomplished by using a cathode follower input stage, followed by an earthed-grid pre-amplifier; the attenuator network is placed between these two valves—V8 and V9 in Fig. 5.

The valves V8 and V9 cannot be selected without careful consideration of various factors. They should ideally require the same bias and take the same anode current. In both, inter-electrode capacitances should be small, and the inductance of leads ought to be small also. Mutual conductance must be high, and anode current cannot be too small or the voltage output may be restricted unduly. In the case of V8 a triode can be used as the anode is earthed to alternating current. V9 might also be a triode, but a pentode has the advantage of having extra screening between cathode and anode because of the extra two grids. Consideration of these and other matters resulted in the choice of the type EC91 for V8 and the type EF91 for V9.

The cathode resistors of these valves have to be low in value to enable the required bandwidth to be obtained. The values are 175Ω and 180Ω respectively; the resistors should be selected to these values within 1% and should be of adequate power rating and preferably of the composition type.

Calculation of Attenuator Resistance

Fig. 6 shows the input circuit in simplified form, and it will be assumed that R1, R2, and R4 are small compared with the anode slope resistance of the associated valves. The method of calculating the values of attenuator resistance will be given, in case the constructor wishes to provide for attenuation ratios different from those used in the prototype oscilloscope.

Impedance at A-A, looking into R1, is $1/gm_1$ in parallel with R1 ....(1)
Impedance at B-B, looking into R2, is similarly $1/gm_2$ in parallel with R2......(2)

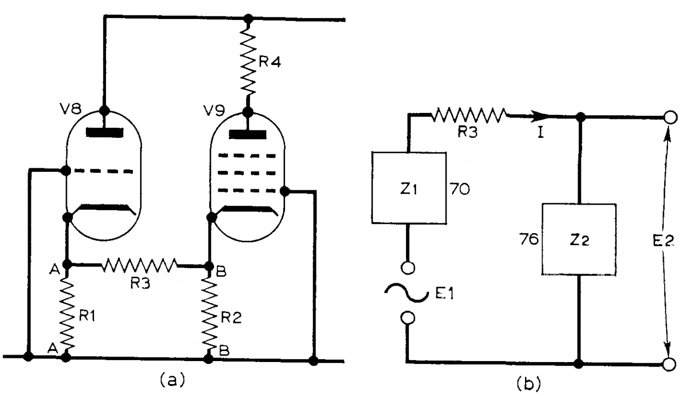


Fig. 6. Analysis of the attenuator.

In (1) above, impedance $Z1 = 70\Omega$ (EC91)
In (2) above, impedance $Z2 = 76\Omega$ (EF91)

Because of an input signal at the grid of V8, a voltage is generated in the circuit formed by Z1, R3 and Z2 (see Fig. 6b). Let this voltage be E1. This causes a current I to flow in the circuit—note, this refers only to the a.c. flowing.

$$\text{Current in the circuit } = I = \frac{E1}{Z1 + Z2 + R3}$$

The output voltage E2 in the figure is the quantity I.Z2 (Ohm's law) and so

$$E2 = \frac{Z2.\ E1}{Z1 + Z2 + R3}$$

$$\text{Attenuation} = \frac{E2}{E1} = \frac{Z2}{Z1 + Z2 + R3}$$

If E2/E1 is replaced by 1/n (n being regarded as the number of "times" the signal is to be divided by)—the expression becomes

$$n = \frac{Z1 + Z2 + R3}{Z2}$$

which simplifies readily to

$$R3 = Z2(n-1) \ - Z1$$

R3 can now be computed for any degree of attenuation with any pair of cathode impedances desired.

Using the values of Z1 and Z2 already found, the following values of attenuation are found:

Attenuation	R3	Attenuation	R3
$\frac{1}{2}$	6Ω	$\frac{1}{16}$	1070Ω
$\frac{1}{4}$	158Ω	$\frac{1}{32}$	2286Ω
$\frac{1}{8}$	462Ω		

To keep the actual values of resistance low in value, a series connection is adopted, and thus the resistors used in the attenuator are 6Ω, 152Ω, 304Ω, 608Ω and 1216Ω. These, counting in series from the first, total to the required value at each step. It may be noted that all these resistors fall within the frequency-independent band mentioned above. Consequently with this simple switched attenuator no compensating capacitors are required.

It will be seen that the value of attenuation is not calculated for the ratio unity. The exercise is left to the reader, and the result is not surprising; for "nil" attenuation ($n = 1$) the result shows that a *negative* resistance is required. A moment's consideration will show that the maximum transfer of energy form Z1 to Z2 is $\frac{1}{2}$, when impedances are equal. Thus an actual attenuation of $\frac{1}{2}$ is seen as "divide by 1", an attenuation of $\frac{1}{4}$ as "divide by 2", and so on. The latter are the values actually marked on the front panel of the instrument.

In practice the 6Ω resistor is included with the 152Ω resistor without appreciable loss of accuracy, and a direct connection used for the minimum attenuation position of the switch. Referring to Fig. 5 the values of resistors R34, R35, R36 and R37 are thus 158Ω, 304Ω, 608Ω and 1216Ω respectively. These are quite close to the nominal preferred values of 150Ω, 330Ω, 600Ω (or 680Ω) and 1·2kΩ, and may be selected from stock. They should be of the miniature $\frac{1}{4}$W type, of carbon composition, and in soldering on to the switch S4 (in series) they should be stood away from the wafer by $\frac{1}{4}$in leads. The leads to the attenuator should be as short as possible.

Maximum Input Voltage

When the attenuator is in its minimum-attenuation position, V8 and V9 act as a long-tailed pair and there is no current feedback. The maximum voltage to V8 grid without overloading is thus the grid base of the valve, namely about 1·5 to 2V. A full vertical scan of the cathode-ray tube is however obtained with this voltage input, and so any larger input requires the attenuator to be switched to a higher position.

In the maximum-attenuation position (÷16) the input voltage acceptable is about 6V peak-to-peak for undistorted vertical deflection, but higher input is permissible if only the positive peaks are to be observed. For greater than 6V input however it is better to use an external attenuator suitably compensated, or the valves V8 and V9 may be replaced by valves with a longer grid base. About 36V input could probably be accepted by a pair of EL90 pentodes, but at the cost of an extra 60 to 70mA of h.t. current—a not very economical proposition!

The valve V9, an EF91, with screen by-passed to earth is not acting as a pentode in this circuit but as a triode. It might be wondered why the EF91 is used rather than another EC91. The reason has already been mentioned, namely the need for the best possible screening between cathode and anode which the extra two earthed grids give. The main amplifier is connected to the anode of V9, and the smallest unwanted feedback could give rise to the most disastrous effects, either in phase distortion or downright instability.

Short Leads Vital

It is most important to ensure that the anode of V8 is bypassed to earth by the 16μF capacitor C25, and decoupled by R32, *right at the valve socket and by the shortest possible leads.* During development the oscilloscope was tested with square waves of very short rise time (about 50ns), and a most conspicuous "ringing" was observed, at a frequency of about 12Mc/s, at the beginning of each horizontal part of the trace.

This effect persisted even when pulses of much longer rise-time were used, though the effect decreased with increase of rise-time. Eventually it was found that the ringing was due to the leads to V8 anode being an inch or

two in length, and relocating the decoupling components effected an immediate cure. A further practical point to be watched is that the input socket to the Y-amplifier cannot rely upon earthing through the front panel of the instrument. A stout braid lead must connect the "outer" of the coaxial socket to the chassis earthing point of the V8 cathode resistor, and an equally stout braid is used for the connection between the "inner" of the socket and V8 grid. Thus inductive effects are minimised. A suitable braid is the outer covering of $\frac{1}{4}$in coaxial cable. Alternatively, a strip of copper foil $\frac{1}{2}$in wide might be used.

Problems of Wide-band Amplification

The necessary Y-amplifier bandwidth of 20Mc/s can be achieved well enough by the use of conventional amplifier techniques. So can the gain needed. However, when an output of some 80V to 100V is needed there are difficulties. A valve such as the 6CH6 with a small anode resistor would provide the necessary output, but at the expense of a large anode current. Also, in order to use a conventional amplifier of cascaded RC circuits the bandwidth of each would have to be increased because of the well-known bandwidth-narrowing effect. This would lead to the use either of compensated couplings, with the attendant problem of overshoot, or of further reduced anode load resistances which by decreasing gain would involve the requirement of extra stages of amplification—possibly with cathode follower stages between the main amplifier valves. Not only is a large number of valves needed but such difficulties as probable unwanted feedback have to be faced. By no means a small problem would be the dissipation of the extra heat developed.

The "Distributed" Amplifier

In order to provide a readily-constructed and "tame" amplifier of high gain and relative freedom from any tendency to instability it was decided to use a "distributed" amplifier. This type of amplifier has in addition the advantage that nearly all the h.t. current supplied contributes directly to the provision of output voltage for Y-deflection at the cathode-ray tube and the heat dissipation problem is much eased.

The pre-amplifier V9, already referred to, has as its anode load a 910Ω resistor R38, and since this valve has to drive only a very limited capacitive load the bandwidth resulting is great. The gain is not about six, as might be expected from considerations of load and mutual conductance, since the cathode resistor is not bypassed—necessarily so, since the input voltage to V9 is developed across it. Actually the gain is approximately 3 for this stage and the anode load resistor could be increased except for the fact that it has to feed the grid line of the distributed amplifier, whose impedance is 910Ω. As will be seen, this stage causes no bandwidth limitations.

The distributed amplifier consists of three EF184 valves. Their mutual conductance is about 15mA/V and since they are effectively in parallel as regards gain, the low-frequency gain is about 100 with an anode load of 2·2kΩ. At high frequencies the gain is the same, of course, because although the valves are in parallel as regards gain they are not in parallel as regards inter-electrode and other capacitances.

Principles of Operation

Referring to Fig. 5, imagine a sudden change of voltage occurs at the anode of V9. This is impressed on the line of inductors L10, L11 and L12 with which are associated the grid-cathode capacitances of valves V10, V11 and V12. The impulse does not travel along this line with the speed of light but at a speed determined by the values of inductance and capacitance in the grid line. As the impulse passes each grid the valve generates a corresponding but larger impulse at its anode.

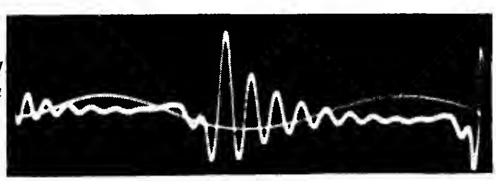
Consider the anode pulse at V10; this has two paths open to it, "left" and "right", and if the impedance of the anode line of inductors is the same as the anode load resistor half the pulse travels in each direction. The pulse going to the "left" is absorbed in the resistance R42, while the rest of the pulse begins to move to the right along the anode line of inductors L5, L6, L7 and L8. If these have been correctly proportioned with the anode capacitances the pulse will reach the anode of V11 just as the grid pulse arrives at the grid, and this process is repeated all down the line. The result is that the travelling pulse is augmented at each anode.

The process may be visualised as a pulse travelling along the anode line and receiving a good push as it passes each anode. Eventually it arrives at the cathode-ray tube deflector plate, where it causes the appropriate deflection and is then *reflected back along the line*, travelling back to R42 where it is absorbed. R42 must match the anode line, or else the pulse will be reflected again and eventually reach the deflector plate once more—mixed up with subsequent pulses generated. This would cause very great interference with the proper functioning of the amplifier, as might be expected.

Line Termination

The valves are seen to be *not* in parallel, and their inter-electrode capacitances do not add up to reduce the bandwidth of the amplifier. They may be considered as being in "delayed parallel". Fig. 7 shows the way pulses are

Fig. 7. Reflection of a pulse in coaxial cable which is open-circuited at both ends.



actually reflected back and forth along a piece of coaxial cable open-circuited at both ends, and this may underline the need for good matching between the anode line and the anode load resistor.

The pulse travelling down the grid line must *not* be reflected at all however, since if it were it would excite the grids in reverse order and cause spurious signals in the anode line.

To prevent reflection the grid line must be terminated in R46, the correct value of resistance, at its "far" end. The small inductors and capacitors L4, L9 and L13, TC1, TC2 and TC3 are added to the delay lines in order to provide that the terminating resistors behave in a frequency-independent fashion at the higher frequencies covered by the amplifier.

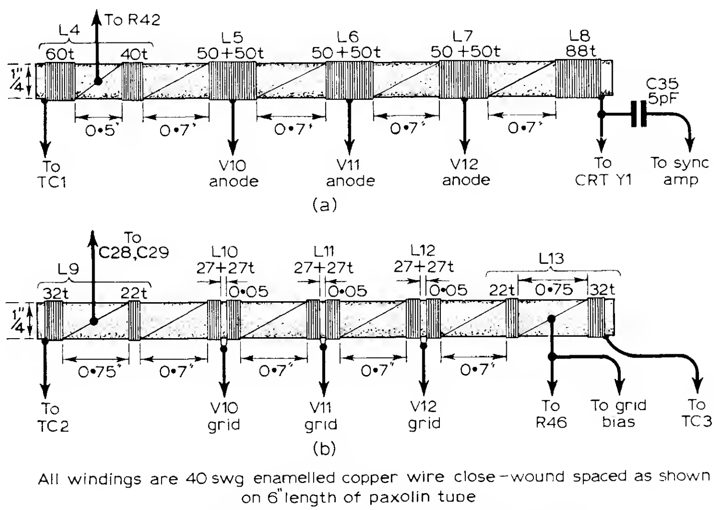


Fig. 8. Winding details of the anode and grid lines.

Winding details for the anode and grid lines are given in Fig. 8, and these must be adhered to strictly if the best results are to be obtained. It should be noted that Fig. 5 shows the anode and grid lines diagrammatically only, and Fig. 8 must be consulted for details of winding and for connections.

Constructing the Delay Lines

The delay lines should be constructed so as to be self-supporting in the circuit. This may be done readily by using fairly heavy-gauge wire for the tappings. The terminal is made by drilling a small hole through the paxolin tube near the end of the winding. A length of 18 s.w.g. tinned copper wire is forced through the hole, emerging on the opposite side of the tube, the end is bent over through 180° and pushed into a further small hole drilled

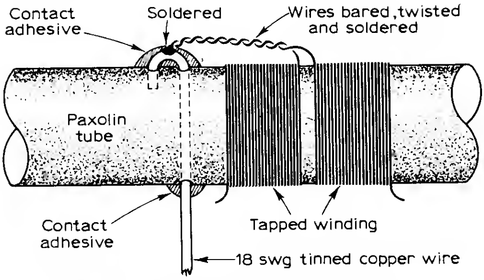


Fig. 9. Termination of the delay-line winding.

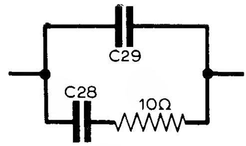


Fig. 10. Additional C and R to eliminate resonance in a coupling capacitor.

in the tube. The winding tapping is then soldered to the terminal and the joint is covered with a good blob of contact adhesive to make all firm. Fig. 9 shows this method of construction in detail.

The terminal wires are then cut all to the same length, about 1·5–2in. These leads can then be soldered direct to the tags of the valve sockets, leaving the delay line supported in the circuit a little over 1·5in above the metal chassis.

The trimming capacitors TC1, TC2 and TC3 are of the Philips beehive type, 2–10pF, and are soldered on to tags fixed to the chassis at convenient points.

Associated Capacitors

It will be noted that C28 and C29 are in parallel, as also are C33 and C34. The reason for this arrangement is that at the higher frequencies the larger-value capacitors may have an inductive reactance. They are paralleled by the smaller mica or ceramic capacitors so as to provide proper coupling. If desired, a 10Ω resistor may be added in series with the larger capacitor; this is only necessary if because of the physical characteristics of the larger capacitor there is any tendency for the paralleled capacitors to behave as a tuned circuit and to resonate at or near some frequency in the pass-band of the amplifier. If this happens it will be shown up as "ringing" on fast pulses or as a kink in the response curve of the amplifier. C29 should be of the smallest physical size practicable in any case, to minimise external capacitance to earth. In the prototype it was not found necessary to include these small resistors, but it will do no harm to include them and may be advisable. Fig. 10 shows the connections.

In order to make the coupling between the distributed amplifier and the pre-amplifier independent of frequency the values of C28 and C33 should be matched as accurately as possible, preferably to 1%; the same applies to C29 and C34.

The screen decoupling capacitors C27, C30, C31 and C32 must be wired as closely as possible to the valveholders, with the shortest possible leads. In

addition, if the length of h.t. lead between the power supply and the amplifier exceeds a few inches an additional 16μF capacitor should be used to decouple the amplifier, and should be placed as close as possible to the tag-strip carrying the service.

The grid bias supply for the distributed amplifier is obtained from the 6·3V heater supply as shown in Fig. 11. This d.c. supply is rectified by D1 and smoothed by C36, C37 and VR6. The potentiometer VR6 may be of wire-wound television type. The value of R48 is fairly critical for the proper functioning of the amplifier over long periods of continuous use. R47 limits diode current on first switching on, and must not be omitted.

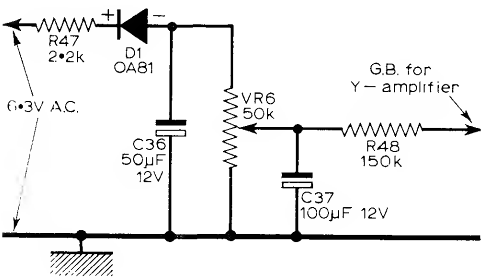


Fig. 11. Grid-bias supplies for the Y-amplifier.

Mains Transformer Construction

If a mains transformer T1 of the type specified in the components list (page 369) is hard to obtain, a suitable component may be wound by hand. The core area should be between $2\frac{1}{2}$sq in and 3sq in; the window size for $1\frac{1}{2}$in E and I laminations would be $2\frac{1}{4}$in $\times$ $\frac{3}{4}$in. This would wind at $4\frac{1}{2}$turns/volt, giving a primary winding of 1080 turns of 26 s.w.g. enamelled copper wire, tapped at 990 and 900 for mains adjustment. Allowing $\frac{1}{8}$in for the cheeks of the bobbin this would amount to ten layers, insulated by thin waxed paper, and the length of this winding would then be 0·2in.

The 300V secondary would require 1350 turns of 34 s.w.g. gauge wire, winding at 100turns/inch. This gives seven layers which with insulation as for the primary would be $\frac{1}{8}$in high, leaving $\frac{1}{4}$in in the window space for the remaining high voltage and heater windings.

The 900V winding could consist of 4050 turns of 40 s.w.g. wire, in 11 layers, amounting to $\frac{1}{10}$in at most with insulation, but this winding must be very well insulated from the other windings by Empire tape.

The secondary heater windings could be wound side by side in the remaining space. The first, of 30 turns of 20 s.w.g. wire (6·3V for the valve heaters) would occupy 1·2in, while the winding for the cathode-ray tube heater would be 30 turns, tapped at 20 turns for 4V, of 22 s.w.g. gauge wire, this would fit nicely into the rest of the winding—about 1in—with plenty of room for insulation.

The above specification is a little tight for hand winding, but by choosing

a larger size of laminations the window space problem will be eased. The same number of turns/volt can be used for the larger transformer, but if a larger number of laminations is selected to give a rectangular rather than a square centre leg, the turns/volt may be reduced, shortening the windings. This cannot be carried too far however.

If, as is quite allowable, an old mains transformer is stripped down to provide laminations, the turns/volt can be ascertained from the windings removed.

The primary, if of sufficiently heavy gauge, can be left in place and the new windings added on top. A transformer of generous proportions should be selected to give a little elbow-room in winding.

Auxiliary Heater Transformer

The auxiliary heater transformer T2 should be readily available from normal suppliers. Two suitable types made by Elstone (MT/LT2) and Osmabet have current ratings of 3A, and will consequently provide a good margin of safety. The physical dimensions of T2 are also of some importance, since the space available for this component is not excessive.

T2 should be wired up to supply the heaters of the Y-amplifier valves V8–V12; it should also be connected to the switch S2 so that it supplies either V7 or the calibration circuit depending upon the setting of the switch.

Chassis Assembly Details

The upper deck will be mounted about 6in above the lower chassis. If the front panel of the instrument and the support carrying the socket of the cathode-ray tube are fixed to the lower chassis, they form a good means of mounting the upper chassis in place. The structure will not be very stable, and will need some kind of bracing; but when the cover is fitted and screwed into position a very strong and rigid assembly will result. However, before the cover is put in place it will be necessary to carry out some or all of the setting-up procedure, and the aim should be to provide enough bracing for reasonable stability without involving screws or bolts which project outside the limits of the bottom chassis. Countersunk screws provide a good solution to this problem.

Individual constructors may have their own ideas concerning the physical arrangement and fabrication of the metal-work. Providing the various points already stressed with regard to critical components and wiring are observed, there is of course no reason why the main structure should be built in a somewhat different form to that employed in the original design.

However, for those who wish to follow the mechanical design of the prototype, dimensioned drawings are included of the several pieces of metal that collectively make up the oscilloscope assembly.

Drilling and Bending Details

In Fig. 12 are details of the drilling and bending for the front panel. It should be noted that if the recommended $\frac{1}{16}$in aluminium is used a bending radius of $\frac{1}{16}$in must be allowed for at each dotted line. For example, the finished width of the front panel is 7in, but it will be seen that in Fig. 12 a dimension of $6\frac{7}{8}$in is indicated between the dotted lines. The missing $\frac{1}{8}$in will be contributed from the two side flanges ($\frac{1}{16}$in from each) when these are bent.

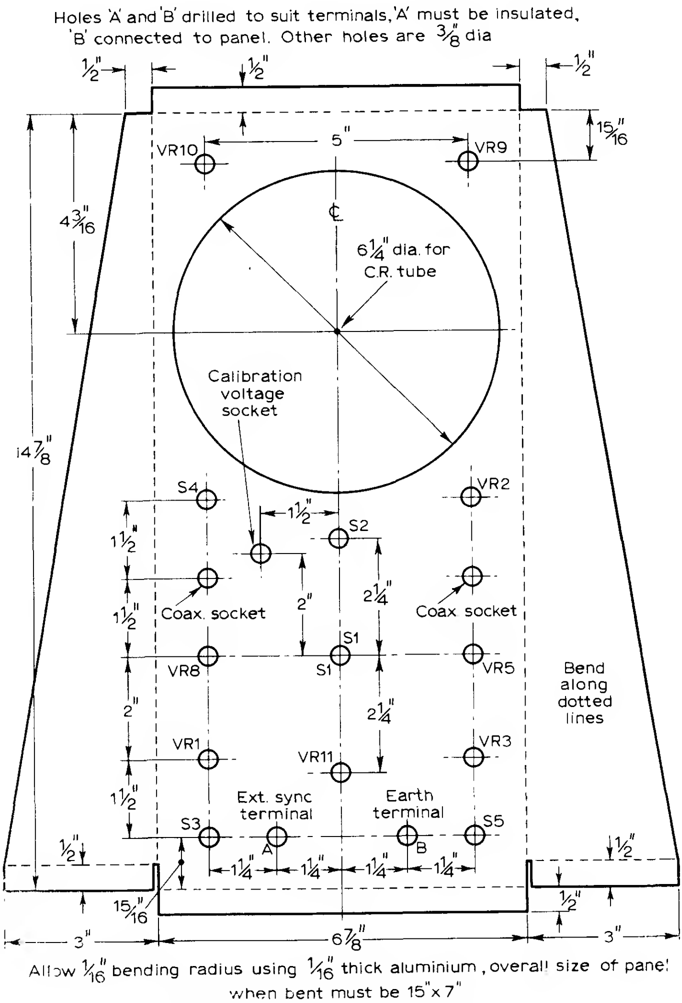
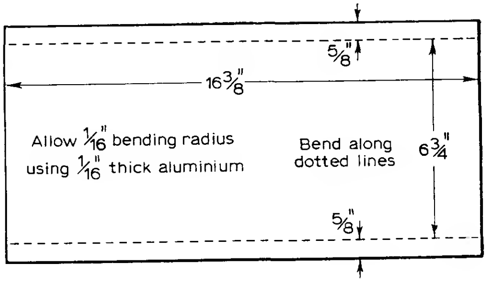


Fig. 12. Drilling details of the front panel.

Fig. 13 (right). *Dimensions of the lower chassis.*



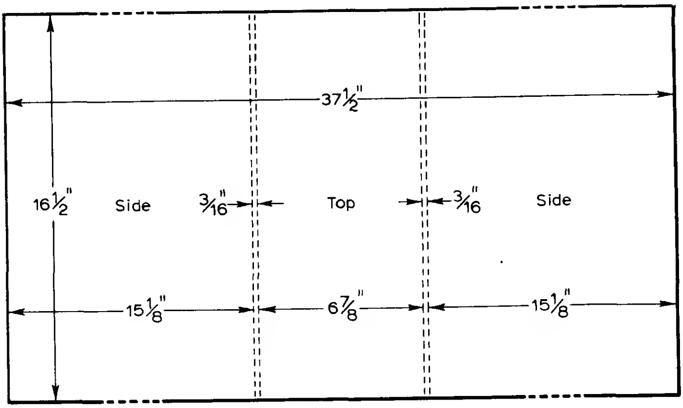


Fig. 14 (left). *The cover of the oscilloscope.*

The bottom chassis is made from a sheet of aluminium according to the details given in Fig. 13. Drilling details for the components are not included, but this should not provide any difficulty, since the actual components may be used as templates.

The cover for the oscilloscope is prepared from a single sheet of aluminium, cut and bent according to the dimensions given in Fig. 14. Since minor discrepancies may arise between the specified dimensions and the actual model produced, this cover should not be prepared until the main assembly has been completed. Then, the measurements given in Fig. 14 can be modified should this prove necessary in order to give a good fit.

Front Panel Wiring

The arrangement of the operating controls, sockets etc., is shown in Fig. 15. The various components should be fitted on the panel and as much of the wiring as is feasible at this stage carried out. It will be necessary to leave most of the flying leads until the final assembly work is reached, as it will then be possible to estimate closely the length required in each case.

Details of the wiring for the X-range switch S1 are given in Fig. 16, while switch S2 is treated in similar detail in Fig. 17.

The wiring of the c.r.t. base and associated components appears in

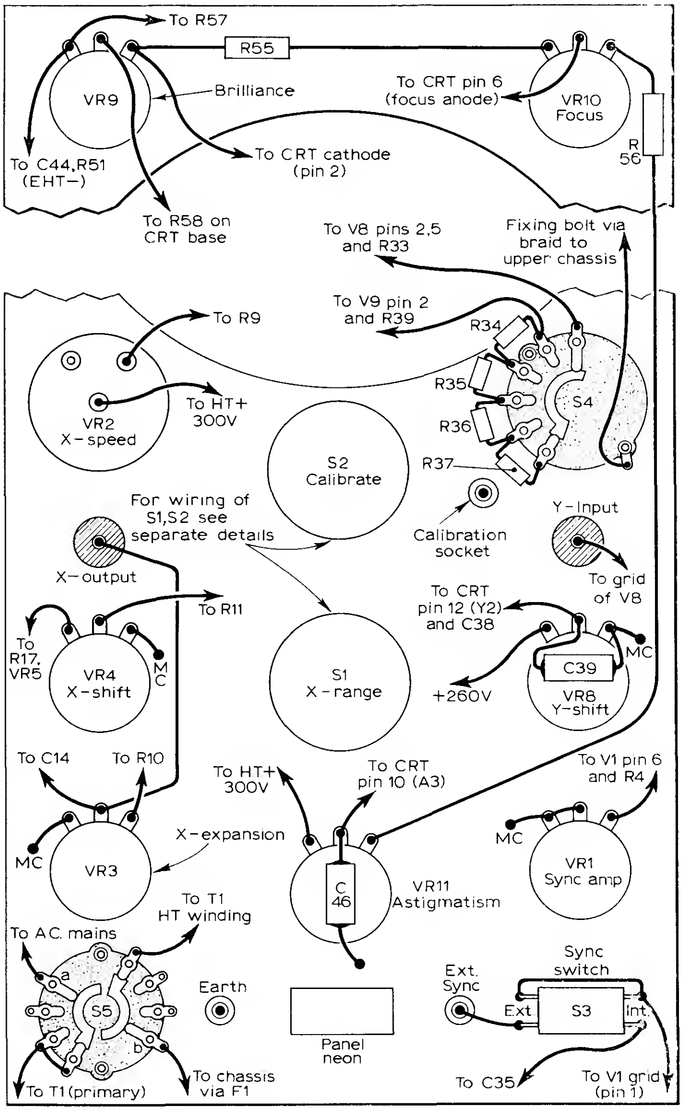


Fig. 15. The layout of the controls, sockets, and wiring behind the front panel.

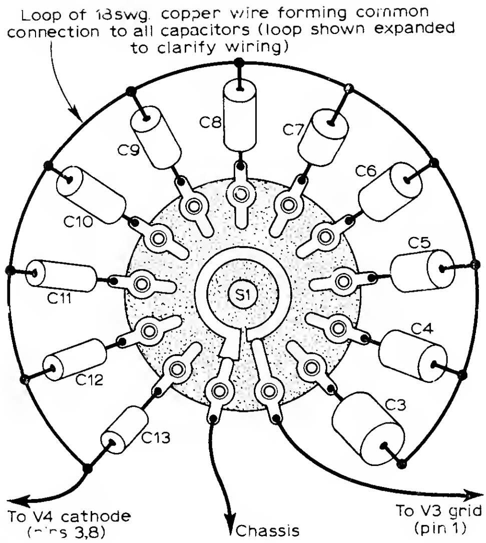


Fig. 16 (right). *Wiring details of S.1*

Fig. 17 (below). *Wiring details of S2.*

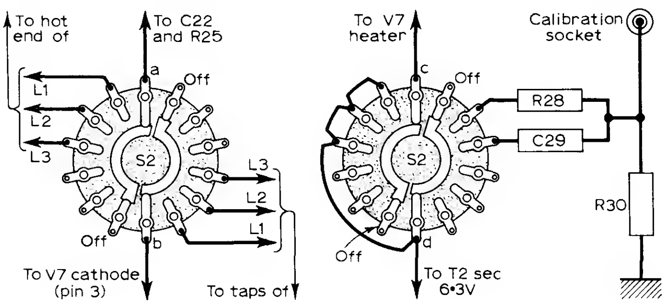


Fig. 18. This wiring should next be completed as far as possible, and then the base and the front panel can be put on one side pending the final assembly of the oscilloscope.

Position of Tube

In order to obtain good visibility with the VCR97 tube it is advisable to make the tube project a little—about $\frac{3}{4}$in—from the front panel; this is because of the curvature of the tube face. A flat-faced tube would be best mounted so that it projected little, if at all. The hole for the VCR97 is cut just 6·25in diameter, and then the tube can be inserted into and withdrawn from its base from the front.

A graticule is a very great convenience in any serious work, especially if it can be removed at will and if it can be rotated to line up the dividing lines

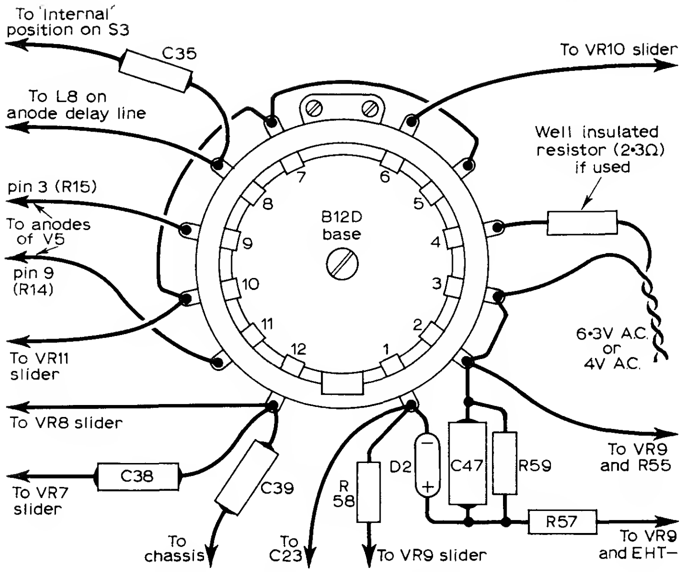


Fig. 18. Wiring details of the base of the c.r.t.

as required. For example, the slope of a wave at any given point can be measured easily in this way.

Aluminium Supports

In the prototype, four strips of aluminium were fixed to the inside edge of the hole in the front panel and bent at right-angles so as to project at right-angles to the panel. These act as supports and guide for the graticule, which is made in the following way.

A tin in good condition is selected from the domestic waste; its diameter should be quite close to 6·25in. The tin is cut short if necessary so that the bottom and about 1in of the side is retained. A hole about 5·5in in diameter is cut in the bottom, and any sharp or jagged edges removed carefully. Light tapping with a small hammer may be useful, against a hard backing, to achieve a smooth and professional looking finish.

Scribing the Graticule

A circle of Perspex is now cut so as to fit snugly inside the tin, and is ruled carefully with a scriber point with lines 1cm apart in two directions at right-angles. A grid of lines 1cm apart is thus produced. These lines may now be inked with black ink, very carefully lest thick lines result. The Perspex disc is now bolted into position in the tin, and the four aluminium strips projecting

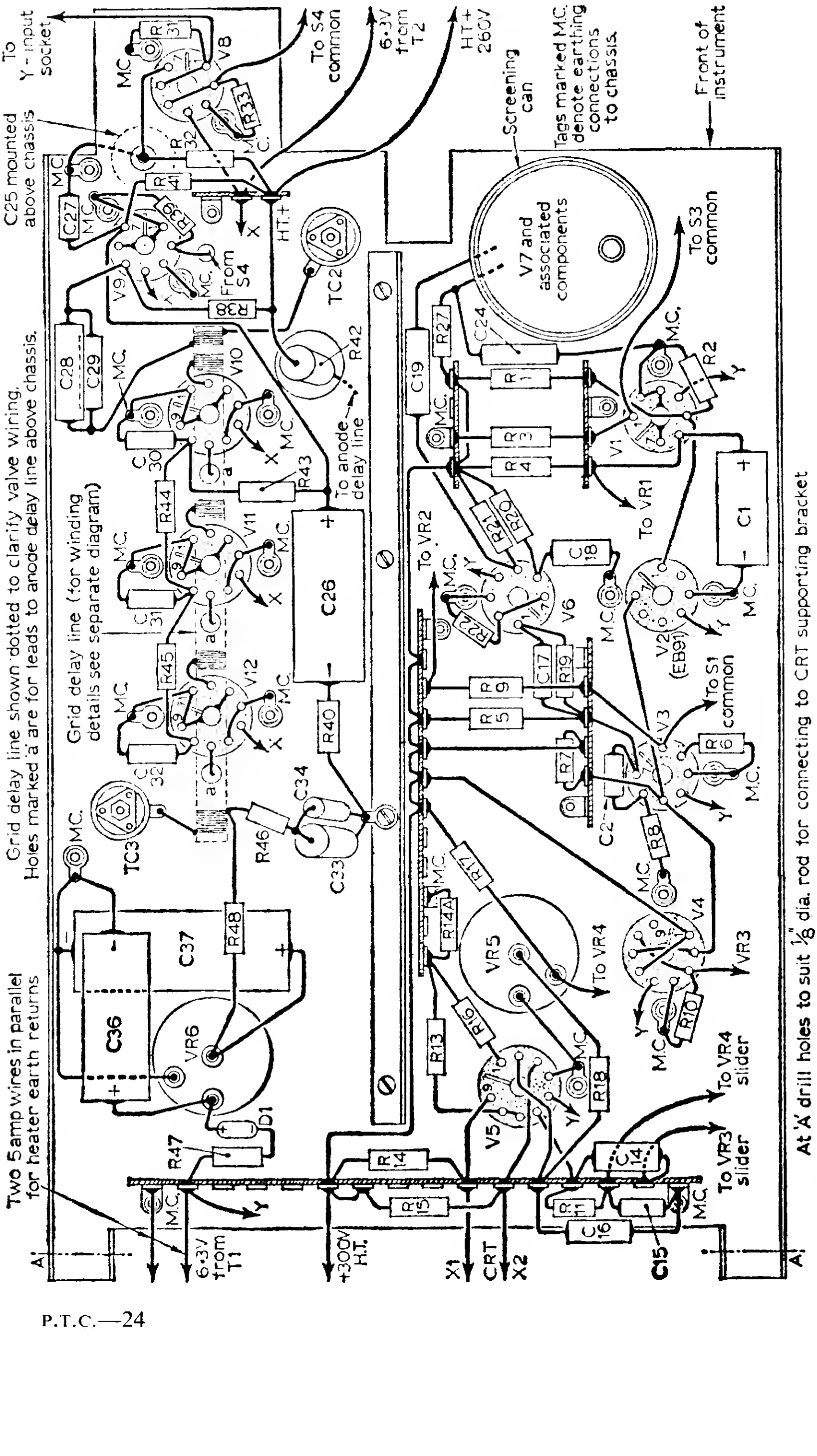


Fig. 19. The above-chassis layout and wiring, with overall dimensions included (further details of the delay-lines are given in Fig. 8).

from the front panel may be bent slightly so that the graticule is held firmly in position. If necessary the aluminium strips may be cut down a little so that the graticule lies as near as possible to the tube face without touching it. It may be that a suitable polish tin will be available, and if it has a rounded rather than a wired edge the appearance will be improved.

Dimensional details of the upper chassis are included in the component layout diagram, Fig. 19.

Since both the timebase circuits and the amplifier carry r.f. currents at high frequency, it is important to prevent cross-coupling. To this end the top chassis, which carries both sets of circuits, is divided down the middle by an aluminium screen some $2\frac{1}{2}$in in height and extending from end to end of the chassis, as nearly as may be.

It may be necessary to cut away small portions of this screen, so that when mounted in position it does not foul the high-voltage capacitors which smooth the e.h.t., or other components on the lower deck.

Wiring-up Details

After assembling the upper and lower chassis in position it will be necessary to make some further connections to the cathode-ray tube socket; the focus, brilliance and astigmatism controls will already have been connected up (see Figs. 15 and 18).

The leads still to be soldered are those to the X and Y plates, and the connections to the network D2, R57, R58 and C47. The coaxial cable to the master oscillator will be readily enough recognised, but the remainder should be identified by tags in some way, or perhaps coded with coloured sleeving.

The settling of the controls on the front panel may be a tricky job, but if reasonable leads have been left where the d.c. controls are concerned the main attention may be given to those carrying r.f. where short leads are essential. For testing purposes these will all have been attached before assembly, and all that remains is to tuck the d.c. leads away neatly inside the space between the chassis, and to shorten the "hot" leads as far as necessary *in situ*.

The lead connecting C35 to S3 cannot be very short, but it should be spaced as clear as possible from any other wires or components to minimise circuit capacitance and to prevent unwanted feedback. *In no circumstances should this lead be enclosed in an earthed braid.*

Good connecting wire for high-voltage components is ordinary p.v.c. covered flex. For additional safety, the soldered points of these components may well be covered with contact adhesive.

The wiring of the power-supply deck is given in Fig. 20.

Screening Cans and Ventilation

Valve screening cans are needed only for the valves V8 and V9. The

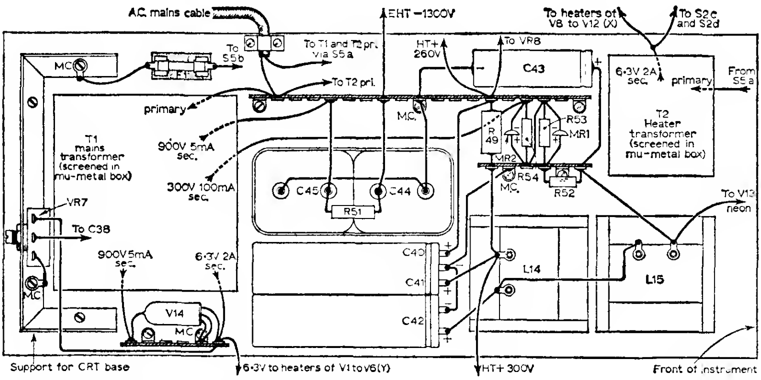


Fig. 20. The wiring of the power-supply deck (see Fig. 13 for the dimensions).

remaining valves are either internally screened or non-critical in this respect, and need good ventilation to dissipate the heat produced.

It has been found that with the instrument case lacquered in black crackle enamel no ventilation louvres are needed, as the temperature rise is relatively small. If desired, a couple of slots each side, top and bottom, will provide all the air circulation needed.

Alternative Valve Types

It is fair to state what alternative valves may be found suitable if the constructor wants to make the most use of his spares box. The timebase generator V3 may be a SP61, V4 may be any good triode, while V5a and V5b may be SP61's connected as triodes, in which case R14 and R15 should be changed to 27kΩ each. V1 and V6 should not be replaced by alternatives, nor should the valves in the Y-amplifier. An ECC81 is by far the best valve for V7, but other double triodes such as the 6SN7 may be found suitable.

Switching On

On switching on the instrument, the "h.t." position of the switch should not be selected until the valves have warmed up; this is readily indicated by the appearance of a green spot on the cathode-ray tube screen, if the brightness is advanced suitably. Since only 1300V e.h.t. is now supplied, the spot is defocused and there is little risk of "burning" the screen.

Allow a few further seconds and select "h.t." when a trace will be displayed. Naturally some five minutes extra should be allowed for "warming up" before any measurements are taken. On switching off, the neon light on the panel will take a few seconds to go out as the smoothing capacitors

discharge. This neon is not used to show whether the instrument is switched on, but to indicate the presence of the h.t. supply.

Setting Up the Oscilloscope

The following procedure is recommended for the final adjustment of all circuits before the cover is put in place.

1. The Distributed Amplifier

(a) Disconnect the h.t. lead to R41 and insert a milliammeter on its 0—100mA range.

(b) Adjust VR6 until the h.t. current is 40mA.

(c) Reconnect R41 to h.t. (temporary joint).

(d) Break the h.t. lead to the whole amplifier. Check that total current is between 62mA and 65mA.

(e) Reconnect the h.t. lead (temporary joint).

(f) Check voltages as follows:

- (i) h.t. supply 255V to 263V.
- (ii) Any EF184 anode pin 185V to 190V.
- (iii) Any EF184 screen pin 175V to 182V.

(g) Allow a half-hour warming-up period; readjust VR6 if necessary and check all through as above.

(h) Allow a further half-hour warming-up period and readjust as above.

2. The Timebase Generator Output Amplifier

(a) Select a medium operating frequency between 5kc/s and 10kc/s; operate shift controls until trace is seen and is central, adjusting brightness as necessary. Focus as accurately as possible, adjusting astigmatism control as needed (this control has a fairly critical setting, but do not worry too much about this at present).

(b) Operate expansion control to obtain a trace about 2in long; no Y-input.

(c) Operating VR4 with one hand and VR5 with the other, adjust until the maximum length of trace is obtained.

(d) Adjust trace length to about 5in and repeat adjustment of VR4 and VR5 until maximum trace length is obtained. Both halves of V5 are now working on the linear part of their characteristic curves, and the trace should be linear with time. Check by applying a sinusoidal voltage to Y-amplifier so as to display about 10 to 20 waves; note that sinewaves are equally spaced. If not quite linear, effect further slight adjustments of VR4 and VR5 for best results. Note that sync will cause a little non-linearity at the very beginning of the trace unless the barest lock is achieved.

(e) If possible check that the adjustment holds on all timebase speed

ranges; no further adjustments should be needed, and if any are so needed suspect a leaking capacitor in the timebase generator or in couplings.

3. Adjusting the Distributed Amplifier

The trimming capacitors TC1, TC2 and TC3 may be put in their minimum-capacitance position for good results, since the required capacitance is about 3pF.

For the best possible results, a pulse generator is needed, and it should generate pulses of rise time less than 40μs. If such is available, apply the pulse to the Y-amplifier, the pre-pulse sync to the sync terminal and select external sync. Adjust sync control to obtain a steady trace.

Adjust the trimmer capacitors in turn until the cleanest possible leading edge and trailing edge are obtained on the pulse. Serious maladjustment will cause loss of steepness on the rising and falling parts of the pulse, and apparent "ringing" on the pulse front, or brightening of the trace at odd spots on the leading or trailing edges. It will not necessarily be possible to remove all of these spots completely unless the pulse generator is guaranteed not to degrade the wave-fronts at all, and it is of course vital to ensure that the pulse generator output is properly matched to its cable and that the cable is properly terminated at the Y-amplifier input. The best adjustment will be easy to find, and if it is found that more than about 5pF is needed, something is wrong somewhere.

If a pulse generator is not available a substitute good enough for this adjustment can be arranged with a thyratron; a suggested circuit shown in Fig. 21.

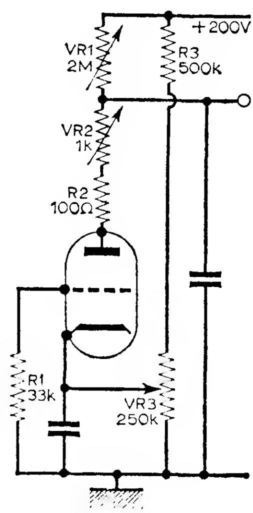


Fig. 21. A substitute for a pulse generator.

VR1 controls the frequency chiefly, and VR3 the amplitude, while VR2 varies the rise time of the flyback, which is the interesting part of the waveform. R2 should not be reduced below 100Ω because even small capacitors give rise to huge currents through the thyratron when flyback occurs and this current needs to be limited to give reasonable valve life. All connections should be kept *very short* to minimise inductive effects, as the frequency spectrum generated on flyback extends well beyond 1000Mc/s.

4. Setting up the Marker Oscillator

The chassis and the screening can will have been drilled for access to the cores of L1 and L2, which are mounted on the same former. The switch S2 should be set to the 1μs range and ten minutes allowed for warming-up.

Select one of the three highest timebase speed ranges by means of S1, and adjust the brightness control so that a row of markers appears on the tube face. Use focus and astigmatism controls as necessary. Apply the output of a signal generator, set to 1Mc/s precisely, to the Y-input, and adjust sync and attenuator controls, and timebase speed control, to obtain a steady trace, with three or more sinewaves displayed. Adjust the core of L1 until a marker dot appears at the same point on each wave. If about 20 waves are displayed it can readily be seen whether the line of marker dots is parallel with the base line of the display.

The same process is repeated, using a lower timebase speed, for display of the 10μs markers. The signal generator input should be at 100kc/s. If the signal generator does not go as low as this, set to 200kc/s and see that a marker dot appears on alternate sine waves.

If no signal generator is available, stray radiation can be picked up on a domestic receiver tuned to 1Mc/s or the long-wave Light Programme as necessary; in the latter case the second harmonic will be picked up. Set to zero beat with the Light Programme.

To adjust the marker oscillator on the third range, 100μs, a calibrated audio signal generator is ideally necessary. If this is not available the method given in the foregoing paragraphs cannot be applied and the procedure should be as follows:

Remove the can surrounding the marker-oscillator assembly to secure access to the end of L3. Display about 30 or 40 waves at 100kc/s and check that every tenth wave is in the centre of the darkest part of the trace, increasing brightness if necessary to limit the dark areas caused by the marker dots; it should be possible to arrange this so that no more than three waves are blacked out. If necessary the tuning capacitance should be varied by substitution to obtain an accurate result. When this adjustment is completed the screening can should be replaced firmly. It may be soldered up if all is well, as it will need to be removed at most infrequent intervals.

Some constructors may feel that 1μs markers are too close together and would like to include a range of 100ns or 200ns markers. The oscillator will

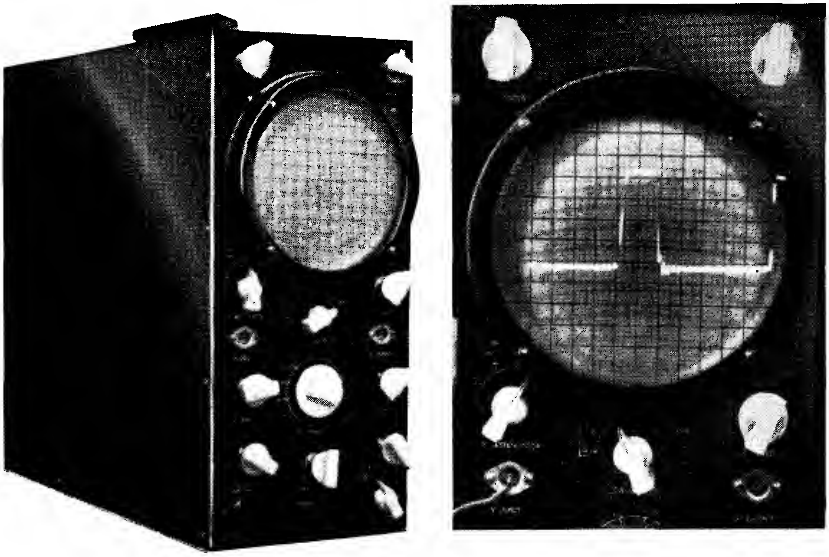


Fig. 22. The finished oscilloscope. (Right) *The face of the instrument, displaying a trace.*

work quite well with a tuning coil arranged to give oscillations at 5Mc/s or 10Mc/s at sufficient amplitude to modulate the cathode-ray tube electron beam. The coil designed for 1Mc/s operation can, if the tuning capacitor is removed and the coil tuned only by strays, be adjusted to operate at 2Mc/s by rotation of the core.

It may be remarked that when the marker oscillator is in circuit, the fly-back trace is automatically blanked. While this has little effect on the lower speed ranges it has the advantage that flyback is not visible even at high brightness settings on the high speed timebase ranges, and the flyback trace does not appear on photographic records at any timebase speed.

The Voltage Calibrator

In order to relate the Y-deflection of the display to voltage input to the Y-amplifier a simple calibrator is provided. This is not switched into the Y-amplifier, as might be thought. The elegance of switching costs too much in input capacitance. Instead, a socket is provided on the front panel, to which either 1V peak to peak or 0·1V peak to peak can be switched at will. This can be applied to the Y-input socket through a short length of wire.

Referring to circuit diagram Fig. 4, it will be seen that a 6·3V r.m.s. supply is divided either by R28 or R30 or R29 and R30 depending upon the setting of switch S2d. 6·3V r.m.s. represents 19·8V peak to peak, and therefore R28 is 1880Ω and R29 is 19,700Ω if R30 is 100Ω. These values should be selected from stock to 1 %. The nearest preferred values are 1·8kΩ and 18kΩ.

The Henlow Wide-band Oscilloscope
Specification

Power Supplies:
200–250V; 40–60c/s; a.c. only; consumption 120W.

Display Tube:
VCR97: electrostatically focused and deflected. 6in diameter screen, medium-persistence green phosphor.
Alternative types: any modern 5in or 6in tube of similar character, not requiring high p.d.a. voltage.
E.H.T.: 1600V, reducing to 1450 for best astigmatism control setting.
Graticule: removable, rotatable, and without illumination. 6in diameter, engraved lines 1cm apart vertically and horizontally.
Spot diameter: 0·5mm maximum at 1600V e.h.t.

Shifts:
X-shift: 100% of maximum deflection.
Y-shift: 500% of maximum deflection. (Both direct-coupled)

Timebase Output:
40V p.p. into 1000Ω shunted by 50pF.

Timebases:
Ranges: less than 30cm/sec to 10cm/μsec in 11 ranges, under unexpanded conditions. With expansion, up to 50cm/μsec. Free running.
Calibration: calibration scales not provided as intensity-modulated markers are available on all ranges.
Linearity: up to 20c/s—10% increasing to 2·5%. 20c/s upwards—3% or better.
Trace expansion: up to 8x normal except on highest ranges, where it decreases to 5x normal expansion from centre of trace. Continuously variable.
Sync: external or internal.
Jitter: not measurable with pulse of 100nsec rise-time.

Y-Amplifier:
Deflection sensitivity: 100mV/cm at 1600V e.h.t.
Impedance: 1MΩ shunted by 12pF.
Response: 12c/s to 20Mc/s.
Attenuator: frequency independent up to 20Mc/s.
Sag: not measurable at 50c/s.
Linearity: 1·5% of full deflection.
Gain control: stepped attenuator giving ÷1, 2, 4, 8, 16.
Accuracy: 1%
Compensation: not necessary.

Mains Surges:
$\left.\begin{array}{l}\text{Y-deflection—2cm}\\ \text{X-deflection—0·3cm}\end{array}\right\}$ for 5% sudden change.

Valve and Semiconductor Complement
Sync amplifier—EF95.
Timebase generator—EF91, EA50 or ½ EB91, ECC82.
Sweep amplifier—ECC84.
Pulse amplifier—EF95.
Marker oscillator—ECC81.
Y-input—EC91.
Pre-amplifier—EF91.
Distributed amplifier—EF184 × 3.
E.H.T. rectifier—EY51.
Bias supply and marker clamp OA81 × 2.
Power supplies: silicon diodes 500mA, 1000V p.i.v. types (2).
Neon indicator—any miniature panel-mounting type.
Limiting fuse—motor side-lamp 6V 0·5A.
Cathode-ray tube—VCR97.

Dimensions:
16½in × 7in × 15in (17½in deep over projecting knobs, 16¼in high over carrying handle).

Weight
19lb.

Cooling:
Natural.

List of Components

Resistors:

R1	47kΩ
R2	47kΩ
R3	1MΩ
R4	100kΩ
R5	100kΩ
R6	220Ω $\frac{1}{2}$W
R7	100kΩ
R8	39kΩ
R9	330kΩ
R10	39kΩ
R11	1·5MΩ
R12	—
R13	2·2kΩ
R14	47kΩ
R14A	6·8kΩ 1W
R15	47kΩ
R16	2·2kΩ
R17	470kΩ
R18	100kΩ
R19	3·3MΩ
R20	120kΩ
R21	100kΩ
R22	220Ω $\frac{1}{2}$W
R23	1MΩ
R24	2·2kΩ
R25	100kΩ
R26	5kΩ
R27	22kΩ $\frac{1}{2}$W
R28, R29	See text
R30	100Ω
R31	1MΩ
R32	100Ω
R33	175Ω $\frac{1}{2}$W
R34, R35, R36, R37	See text
R38	910Ω $\frac{1}{2}$W 5%
R39	180Ω $\frac{1}{2}$W
R40	10Ω
R41	5kΩ 1W
R42	2·2kΩ 3W 5% h.s.
R43	10Ω
R44	10Ω
R45	10Ω
R46	910Ω $\frac{1}{2}$W 5%
R47	2·2kΩ
R48	150kΩ
R49	1kΩ 10W
R50	1MΩ
R51	470kΩ
R52	47Ω 2W
R53	560kΩ
R54	560kΩ
R55	470kΩ
R56	3·3MΩ
R57	1M
R58	250kΩ
R59	560kΩ

All 10% $\frac{1}{4}$W except where otherwise indicated.

Variable Resistors:

VR1	100kΩ carbon potentiometer
VR2	2·5MΩ carbon potentiometer
VR3	50kΩ carbon potentiometer
VR4	1MΩ carbon potentiometer
VR5	1MΩ carbon potentiometer
VR6	50kΩ carbon potentiometer
VR7	22kΩ wire-wound potentiometer
VR8	1MΩ carbon potentiometer
VR9	250kΩ carbon potentiometer
VR10	1MΩ carbon potentiometer
VR11	2MΩ carbon potentiometer

All the above are linear law.

Capacitors:

C1	8μF electrolytic 250V
C2	5pF silver mica
C3	1μF paper 150V
C4	0·5μF paper 200V
C5	0·15μF paper 200V
C6	0·05μF paper 200V
C7	0·01μF paper 200V
C8	2000pF silver mica
C9	500pF silver mica
C10	200pF silver mica
C11	50pF silver mica
C12	22pF silver mica
C13	(strays)
C14	0·5μF paper 200V
C15	5000pF silver mica
C16	0·5μF paper 200V
C17	5pF silver mica
C18	0·5μF paper 300V
C19	0·25μF paper 250V
C20	1500pF silver mica
C21	1500pF silver mica
C22	50pF silver mica
C23	30pF silver mica 2500V
C24	0·25μF paper 250V
C25	16μF electrolytic 300V

List of Components (*contd.*)

Capacitors (contd.)
- C26 32μF electrolytic 250V
- C27 1500pF silver mica
- C28 0·5μF paper 300V
- C29 1500pF silver mica
- C30 1500pF silver mica
- C31 1500pF silver mica
- C32 1500pF silver mica
- C33 0·5μF paper 300V
- C34 1500pF silver mica
- C35 5pF silver mica
- C36 50μF electrolytic 12V
- C37 100μF electrolytic 12V
- C38 0·01μF paper 300V
- C39 0·1μF paper 300V
- C40 250μF electrolytic 275V
- C41 100μF electrolytic 300V
- C42 100μF electrolytic 300V
- C43 32μF electrolytic 450V
- C44 0·25μF paper 2500V
- C45 0·25μF paper 2500V
- C46 0·01μF paper 300V
- C47 0·1μF paper 150V

Variable Capacitors:
- TC1 2—10pF concentric type trimmer.
- TC2 2—10pF concentric type trimmer.
- TC3 2—10pF concentric type trimmer.

Inductors:
- L1–13 See text.
- L14 Smoothing choke 5H 100mA low resistance.
- L15 Smoothing choke 5H 100mA low resistance.

Transformers:
- T1 Mains transformer. Tapped primary. Secondaries: 0–900V 5mA; 0–300V 100mA; 6·3V 2A; 6·3V (tapped at 4V) 1A.
- T2 Heater transformer. Tapped primary. Secondary: 6·3V 2A.

Switches:
- S1 1-pole, 12-way rotary switch.
- S2 4-pole, 6-way rotary switch.
- S3 1-pole, 2-way toggle switch.
- S4 1-pole, 5-way rotary switch.
- S5 2-pole, 3-way rotary switch.

Valves:
- V1 6AK5
- V2 EA50
- V3 EF91
- V4 ECC82
- V5 ECC84
- V6 6AK5
- V7 ECC81
- V8 EC91
- V9 EF91
- V10 EF184
- V11 EF184
- V12 EF184
- V13 Neon 210—250V
- V14 EY51
- V15 VCR97

Rectifiers:
- MR1, 2 Silicon power rectifier 500mA 1000 p.i.v. E250C50.
- D1 OA81 germanium signal diode.
- D2 OA81 germanium signal diode.

Miscellaneous:

Two coaxial sockets. Two terminals. Valveholders: five B7G; six B9A; one B3G. Lampholder. Panel-mounting holder for neon tube. Base for c.r.t. Perspex sheet 6¼in square.

Index

Practical Television Circuits

NEWNES